# IF FILTER OPTIMIZATION
# FOR CW-FM RADAR

Ergin Kislali

# NAVAL POSTGRADUATE SCHOOL
## Monterey, California

# THESIS

IF FILTER OPTIMIZATION

FOR CW-FM RADAR

by

Ergin Kislali

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A continuous-wave frequency-modulated radar receiver and the IF filter are described.

IF filter responses are investigated from the output of the narrow bandpass filter for various target ranges. The effects of the IF signal pulse width and the duty cycle on the IF filter output including the change of the sidelobe levels out of the filter are examined.

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by

Ergin Kislali
Lieutenant, Turkish Navy
B.S.E.E., Naval Postgraduate School, 1974


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IF filter responses are investigated from the output of the narrow bandpass filter for various target ranges. The effects of the IF signal pulse width and the duty cycle on the IF filter output including the change of the sidelobe levels out of the filter are examined.

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# ACKNOWLEDGEMENT

To my wife Engin.

# I. CW-FM RADAR AND IF FILTER

## A. CW-FM RADAR

The continuous-wave frequency-modulated radar transmitter produces a CW signal of constant amplitude whose frequency varies in a sawtooth fashion as shown in Figure 1. The receiver picks up a part of the echo signal after a delay time of $\tau = 2R/c$. The dashed lines in Figure 1(a) represent the echo signal. If the echo signal is heterodyned with a portion of the transmitted signal, a beat note $f_b$ is obtained as shown in Figure 1(b).

Note that for a given target two different beat frequencies are obtained. The beat frequency obtained during the later part of the transmitted frequency ramp is directly proportional to the target range and is given by

$$f_b = \frac{2R\Delta f}{cT}$$

The block diagram of a CW-FM radar with a sideband superheterodyne receiver is shown in Figure 2. It should be noted that separate antennas have been used for transmission and reception. The IF (intermediate frequency) signal produced by the local oscillator is mixed with a portion of the transmitted signal. The output of the mixer consists of the carrier, sidebands on either side of the carrier, and higher harmonics. A filter selects one of the sidebands as a reference signal and rejects the

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carrier and the other sideband. The sideband filter must
have a large enough bandwidth to pass the modulation, but
not the carrier or the other sideband. When an echo signal
is present and mixed with the output of the above filter
in the receiver mixer, one obtains a signal the frequency
of which is $f_{IF} + f_b$. This signal is amplified and fed
into the balanced detector together with the signal $f_{IF}$
coming from the local oscillator. The output of the detector
contains only the beat frequencies.

$$f_b = \frac{2R}{c} \cdot \frac{\Delta f}{T} \qquad f_b' = \Delta f - f_b$$

The beat frequencies are amplified by a low-frequency
amplifier and then applied to the frequency-measuring circuits.

The radar is normally designed so that $T > 2R_{max}/c$
where $R_{max}$ is the maximum target range of interest. There-
fore the duration of $f_b$ is greater than the duration of $f_b'$.
The later tone is therefore dropped by turning the amplifier
off during the initial portion of the sweep. Therefore,
$f_b'$ will be ignored in the following discussion.

B. DOPPLER EFFECT

When an electromagnetic wave of frequency $f_o$ is trans-
mitted to a moving target, the received echo is shifted in
frequency by an amount $f_d$ according to the Doppler
equation

$$f_d = - \frac{2f_o V_r}{c} = - \frac{2f_o}{c} \cdot \frac{dR}{dt} = - \frac{2V_r}{\lambda}$$

where $V_r$ represents the relative radial component of velocity of the target, C the velocity of propagation and $\lambda = C/f_o$ the wavelength of the transmission. Thus the doppler frequency shift yields information about the velocity of the target. For a moving target we have, then,

$$f_b = \frac{2R\Delta f}{cT} - \frac{2f_o V_r}{c}$$

Solving this equation for target range we have

$$R = f_b \frac{cT}{2\Delta f} + V_r \frac{f_o T}{\Delta f}$$

If $V_r$ is unknown and the range is calculated from the first term in the above expression, there is a range error, $\Delta R$, given by

$$\Delta R = V_r \frac{f_o T}{\Delta f}$$

For surface targets this range error can generally be made small. For example [5], if $f_o$ = 9 GHz, T = 380 microseconds and $\Delta f$ = 12 MHz, the range error is only 16.1 yards for a 100 knot relative radial velocity. In the following discussion it will be assumed that $\Delta R$ is small enough so that it can be ignored.

## C. NARROW BANDPASS IF FILTER

The optimum IF filter in terms of maximum signal-to-noise ratio is matched to the spectrum of the beat frequency in the IF amplifier. Since there is a different beat frequency for each possible target range, the optimum receiver must contain a large number of matched filters, one for each possible target range. For simplicity one IF filter can be used, and the spectrum of each target may then be shifted in sequence to that frequency by varying the frequency of the IF oscillator.

If the received echo signal is a pulse wave of finite duration, $\tau$, and the pulse is repetitive and coherent, then the frequency spectrum of the wave has the shape shown in Figure 3 for the case where $T = 2\tau$. In practice the pulse train is finite, consisting of only one or a very few pulses, and successive pulses are not necessarily coherent. Moreover, there is a slow frequency modulation of the IF signal due to the sweep of the IF oscillator. The spectrum is therefore much more complex than shown in Figure 3 although its envelope is approximately in the SinX/X form shown.

In general, the beat frequency of the echo signal lies somewhere within a relatively wide band of frequencies. Let us assume that such a signal is brought to a bank of narrowband filters, as shown in Figure 5(a), whose combined range of frequencies covers the entire band of the

beat frequencies. Then one of the narrowband filters whose center frequency matches the frequency of that portion of the echo which is due to the target permits the identification and measurement of the target frequency and hence its range.

The bandwidth of an individual filter is wide enough to pass the signal, but not too wide to introduce more noise than need be. These filters intersect at their half-power points and have equal bandwidths as shown in Figure 5(b).

A true matched filter would have the SinX/X frequency response of Figure 3. It is not practical to build such a filter, so a compromise must be reached. A compromise that is often used is for the filter response to be similar to the central lobe of the SinX/X envelope.

## II.   RANGE SIDELOBE REDUCTION

The spectrum of a pulse signal has essentially SinX/X shape, with sidelobes extending on either side of the main-lobe.  The first sidelobe is only 13.2 dB below the peak of the mainlobe, and the near sidelobes fall off at approximately 4 dB per sidelobe interval.  Since these sidelobes may be mistaken for a target at a different range than that of the target, they are called range side-lobes.  The range sidelobe null points are spaced $1/\tau$ hertz apart for the ideal SinX/X waveform.  The range sidelobes of the strong target signals can obscure weaker signals as shown in Figure 6.

CW-FM radars exhibit in general a range ambiguity problem when signals are received from a large and a small target in the near vicinity of each other.  In such cases the small target signals cannot be distinguished.  An IF filter can respond to the range sidelobes of the strong-target signal spectrum.  This spurious response could make it difficult to detect a small target such as a navigational buoy in the vicinity of a large object such as a ship.

The response of the receiver to the undesirable range sidelobes can be reduced by careful selection of the IF filter amplitude response.  However optimizing the filter to minimize the range sidelobes will cause a reduction in output signal-to-noise ratio.  The optimum filter for

maximum output signal-to-noise ratio would be matched to the signal spectrum. Increasing the output signal-to-noise ratio by optimum filtering improves the ability of the receiver to detect the presence of a weak signal. In order to make the target echo signal relatively more detectable, a narrow bandpass filter centered at frequency $f_{IF} + f_b$ should be employed to pass the signal which falls within its passband but rejects all noise outside its passband. The maximum peak output signal-to-noise ratio in terms of the signal energy for the optimum matched filter is given in Refs. 2 and 3.

$$( \frac{S}{N} )_{out} = \frac{2E}{N_o}$$

E is the signal energy, $N_o$ is the noise power density in terms of watts per hertz.

The range sidelobes may be reduced either by amplitude weighting the envelope of the transmitted FM signal or by an amplitude weighting of the receiver filter frequency response. Proper shaping of the envelope of the signals can significantly reduce the sidelobe levels. Figures 7 and 8 [Ref. 4] show that a reduction in sidelobe levels is accomplished by employing the truncated Gaussian and truncated Butterworth signals in an FM chirp (low duty cycle) pulse compression radar. For the amplitude weighting shown range sidelobes are down 35 dB, whereas without weighting they

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would have been down only 13 dB. Figure 9 [Ref. 3] shows
us typical chirp radar receiver output waveforms resulting
from use of the frequency weighting techniques in which the
response far from the carrier falls off more rapidly than
does matched filter response. These waveforms are obtained
from the weighting function

$$W(\omega) = k + (1 - k)\cos^n \frac{\pi(\omega - \omega_o)}{\Delta\omega}$$

Because of its effect on the shape of this frequency
response, k is called the pedestal function. The weighting
function has two effects on the system output. First, the
amplitude of range sidelobes is reduced by an amount
depending on the values of k and n. Second, since the
combined filter is no longer a matched filter, the signal-
to-noise ratio is slightly reduced and the pulse width
increased. Increased width and the sidelobe reduction
are shown in Figure 9 for several values of n and k.

In the present study an attempt is made to optimize the
relative widths of the spectrum and the filter to reduce
the amplitude of the range sidelobes in the case of the
FM-CW radar. Here, the IF spectrum is quite different than
for the FM-CW radar. The question is, can range sidelobes
be reduced in this case by the same or different techniques?

## III.   EXPERIMENTAL RESULTS

In order to determine the effect of the relationship between filter bandwidth and the IF pulse width and duty cycle, the experimental set-up shown in Figure 10 was used. The frequency synthesizer generated a CW signal similar to the signal in the IF amplifier of the FM-CW radar.  The pulse modulated IF signal shown in Figure 11 was fed to the input of the filter.  The output of the filter was measured by means of the power meter.  The time response of the filter was displayed on an oscilloscope as shown in Figure 12.  By changing the width and duty cycle of the input signal pulse various filter-output responses were obtained.  The responses obtained by changing the IF frequency in 100 hertz steps to simulate different target ranges are shown in Figures 16 through 30.  Evidently, the pulse width and duty cycle affect the response of the filter and the amplitude of the range sidelobes in the filter output.

A careful examination of the IF filter responses shown in Figures 16 through 30 shows that both the width and the magnitude of the range sidelobes of the IF filter response are considerably affected by the period and duration of the pulse modulated IF signal.  It is also seen that for the conditions of Figure 20 the first order range sidelobes are lowest with respect to the peak.  In this case the

signal spectrum shape almost matches the filter spectrum
shape.  In this particular case (mention parameters-pulse
width, duty cycle), the half-amplitude bandwidth of the
filter happens to almost coincide with the bandwidth of
the main lobe of the signal.  It is also noted from
Figure 20 that the first range sidelobe of the filter out-
put response is 11.1 dB lower than the main lobe and
lower than the sidelobes obtained in all other figures.
This reduction has been achieved, as noted earlier, by
matching the shape of the received wave form spectrum to
the filter spectrum shape.  Thus, the maximum energy is
transferred from the filter when the shape of the central
portion of the signal spectrum matches the filter spectrum
shape.  As a by-product of this matching, the relative
amplitude of the range sidelobes is reduced.

It is also seen from Figures 16 through 30 that the
width of the response, $\delta_f$, is narrowest for the Figure 20
case.  Since

$$f_b = \frac{2R}{c} \cdot \frac{\Delta f}{T}$$

it follows that

$$\delta_{f_b} = \delta_f = \frac{2\Delta R}{c} \cdot \frac{\Delta f}{T}$$

Solving for $\Delta R$, we have an expression for range resolution

$$\Delta R = \frac{cT}{2\Delta f} \delta_f$$

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Thus the conditions of Figure 20, spectrum matched to filter width, also give the best range resolution.

# IV. ANALYTICAL SOLUTION

## A. COMPUTATION OF POWER SPECTRA

### 1. The Fourier Transform

Suppose that a periodic time function $f(t)$, having duration $\tau$, is to be represented by a Fourier series. The form of the series is then given by

$$f(t) = \frac{a_o}{2} + \Sigma[a_n \cos(2\pi nft) + b_n \sin(2\pi nft)] \tag{1}$$

in which the coefficients $a_n$ and $b_n$ are

$$a_n = \frac{2}{T} \int_o^T f(t) \cos(2\pi nft)dt \tag{2}$$

$$b_n = \frac{2}{T} \int_o^T f(t) \sin(2\pi nft)dt \tag{3}$$

and $f = 1/T$ is the fundamental frequency. Non-periodic waves can be analyzed through the use of the Fourier integral which is an extension of the Fourier series. Such a function may be transformed from the time domain to the frequency domain and vice versa by the following transform pair.

$$f(t) = \frac{1}{2\pi} \int_{-\infty}^{+\infty} F(f) \, e^{j2\pi nft} \, df \tag{4}$$

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$$F(f) = \int_{-\infty}^{+\infty} f(t) \ e^{-j2\pi nft} \ dt \tag{5}$$

The function F(f) is called the Fourier transform of f(t),
and f(t) is called the inverse Fourier transform of F(f).
The integral in F(f) analyzes the time function into a
continuous spectrum.

Over a time $0 \leq t \leq \tau$ , the power spectrum of f(t)
will be estimated by squaring the absolute value of the
Fourier transform F(f).  No signals have been taken outside
the interval $0 \leq t \leq \tau$ so that the integration needs to
be performed only over this time interval.

The spectra can be determined by using the Fourier
transform formula for F(f).  Then the Fourier transform
is computed and the spectral power density is obtained as
$F^2(f)$.  Let us illustrate this procedure by applying it to
a sinusoidal pulse of length $\tau$.  Defining f(t) as

$$f(t) = \begin{cases} A \ \mathrm{Sin}(2\pi f_o t) & - \tau/2 \leq t \leq \tau/2 \\ \\ 0 & |t| > \frac{\tau}{2} \end{cases} \tag{6}$$

and inserting Eq. (6) into Eq. (5) one has

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$$F(f) \equiv A \int_{-\tau/2}^{\tau/2} \text{Sin } 2\pi f_o t \; e^{-j2\pi ft} \; dt \qquad (7)$$

Writing $\text{Sin} 2\pi f_o t$ as,

$$\text{Sin } 2\pi f_o t = \frac{e^{j2\pi f_o t} - e^{-j2\pi f_o t}}{2j} \qquad (8)$$

the integral in Eq. (7) takes the form,

$$F(f) = \frac{A}{2j} \int_{-1/2}^{\tau/2} [e^{2\pi(f_o - f)t} - e^{-2\pi(f_o - f)t}] \; dt \qquad (9)$$

After integration, $|F(f)|^2$ becomes

$$|F(f)|^2 = [\frac{A^2 \tau^2}{4}][\frac{\text{Sin}\pi(f_o - f)\tau}{\pi(f_o - f)\tau} + \frac{\text{Sin}\pi(f_o + f)\tau}{\pi(f_o + f)\tau}]^2 \qquad (10)$$

A part of the spectral power density given by Eq. (10) is shown in Figure 4. There is an identical spectrum for negative frequencies.

## 2. Power Transform through a Network

Suppose that a periodic signal f(t) is applied to a filter with a transfer function H(f) and the output $f_o(t)$ obtained from the product of the Fourier series and the transfer function

$$f_o(t) = \sum_{n = -\infty}^{\infty} H(nf_o)F \, e^{j2\pi n f_o t} \tag{11}$$

The power of f(t) is given by (assuming R = 1 ohm)

$$S = \sum_{n = -\infty}^{\infty} |F(f)|^2 \tag{12}$$

while the power of $f_o(t)$, the filter output reduces to

$$S_o = \sum_{n = -\infty}^{n = \infty} |H(nf_o)|^2 \, |F(f)|^2 \tag{13}$$

## B. ANALYTICAL SOLUTION

For the analytical solution an analytic filter characteristic was selected with a 6 dB bandwidth about the same as the experimental filter characteristic.

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Figures 13 and 14 show us the characteristics of the physical and analytic filters, respectively. The transfer function of the two-pole filter is given by Eq. (14),

$$H(f) = \frac{J\omega KQG_m}{4[(\frac{f-f_o}{f})Q]^2 - 1 - K + j4(\frac{f-f_o}{f})Q} \tag{14}$$

Here K is the smoothing factor of the filter characteristic. K was taken to be unity, Gm is a constant scaled to make the maximum gain unity, and $f_o$ = 30 MHz.

The magnitude of the output of the filter was next computed for various frequency offsets, in 200 hertz steps between the center of the filter. This is equivalent to moving the filter across the spectrum as was done experimentally. Convolution was employed for this computation, and the computation was made for four different combinations of pulse width and period. Figure 15 shows the computed responses in each case. It is also seen from Figure 15 that range sidelobe amplitude levels are reduced by matching the mainlobe of the signal to the filter spectrum shape.

## V.   CONCLUSIONS

It has been shown that the amplitudes of the range
sidelobes in the output of the FM-CW radar of Figure 2 are
a function of the relative widths of the main lobe of the
IF spectrum and of the width of the filter passband.
Experimental work employed a four-pole filter while
theoretical work used a two-pole filter.  In both cases
the range sidelobe amplitude proved to be minimal when
the 6 dB filter width matched the 6 dB width of the central
lobe of the signal spectrum.  With the four-pole filter
used experimentally the total signal fell off more rapidly
with change of range than two-pole filter.

It is probable that range sidelobes can also be
reduced by amplitude weighting of the signal.  Time did
not permit an investigation of this.
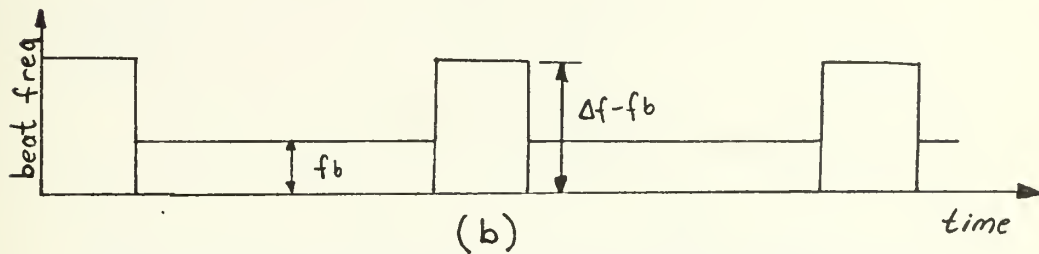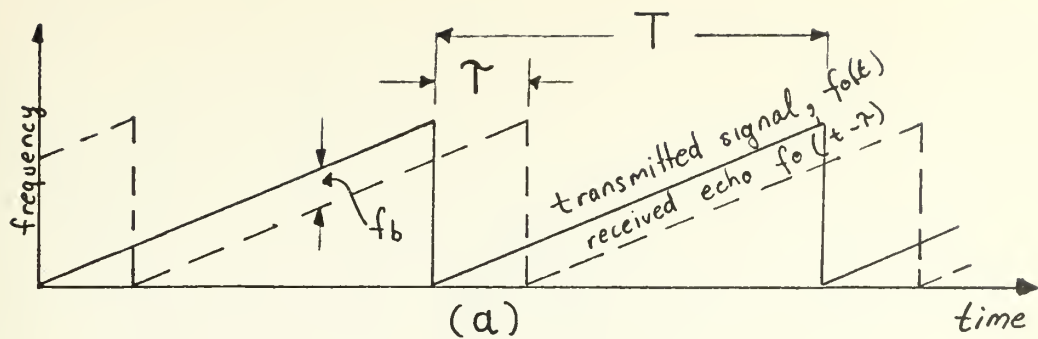
Figure 1 (a) Sawtooth frequency modulation
(b) Beat note
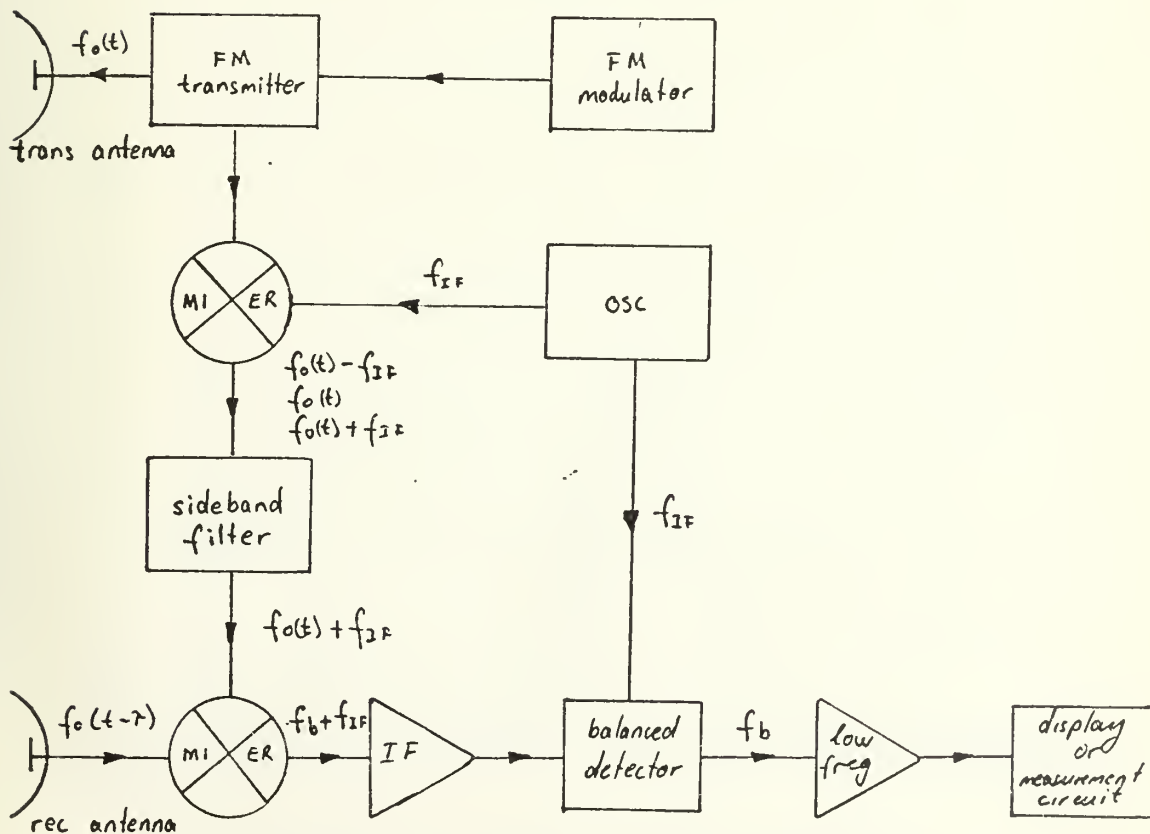


Figure 2   Block diagram of CW-FM radar
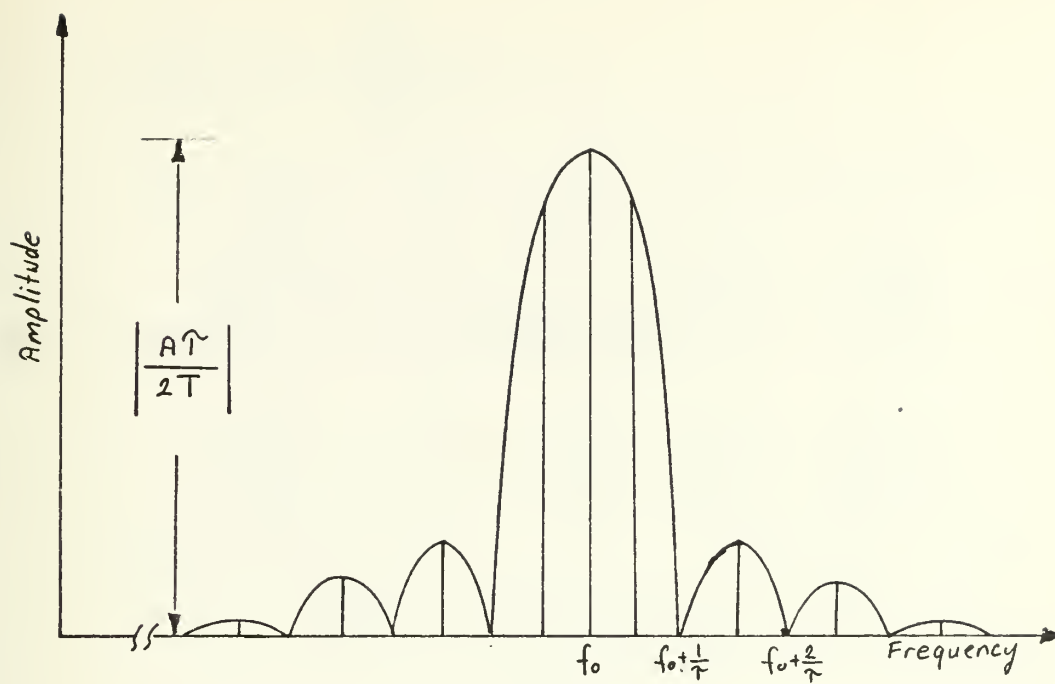
Figure 3    Frequency spectrum of a repetitive
square wave pulse
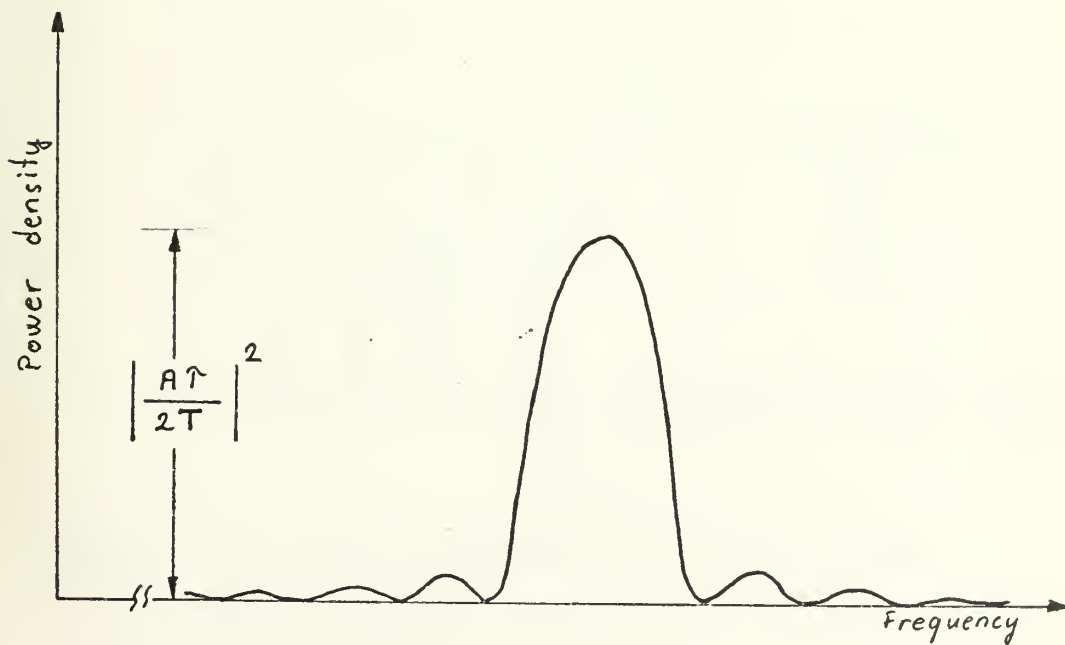


Figure 4    Power spectral density
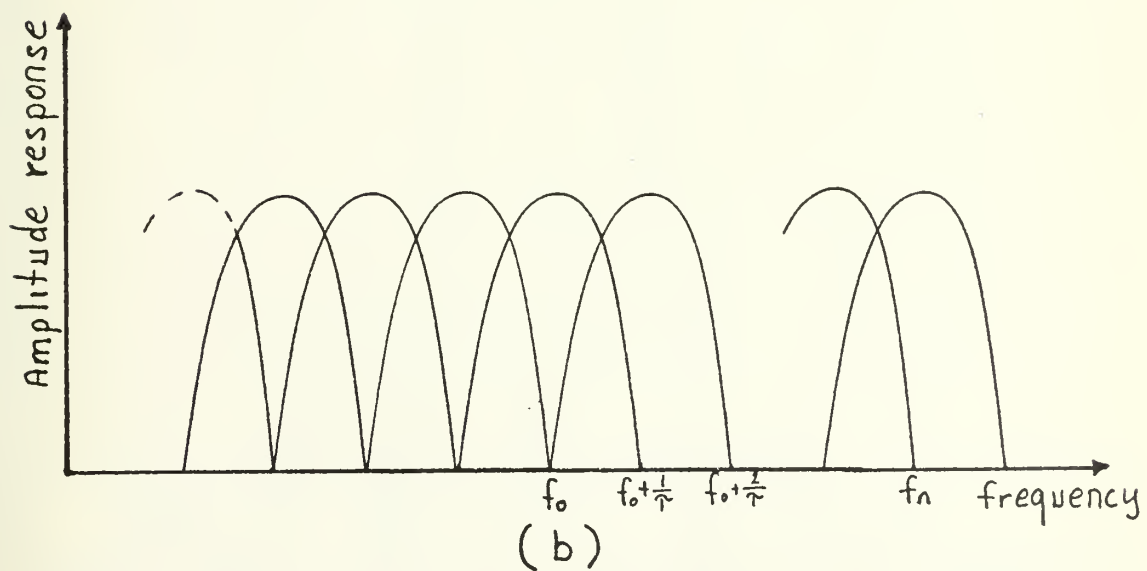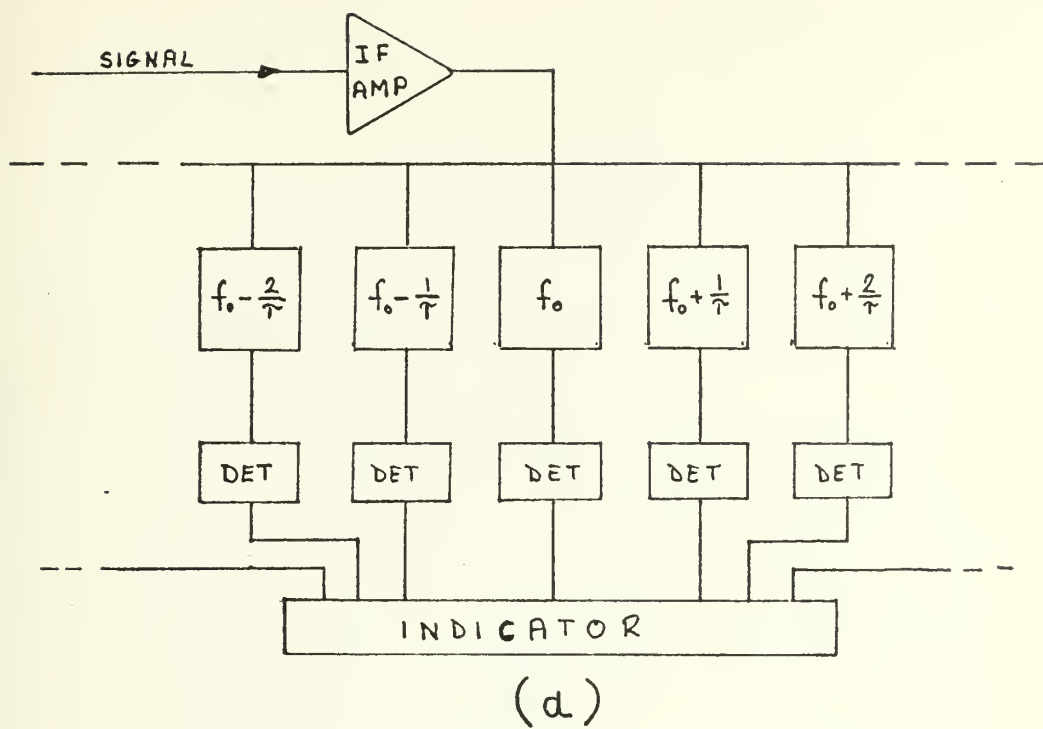
Figure 5 (a)  Block diagram of IF filter bank
         (b)  Frequency response of IF filter bank
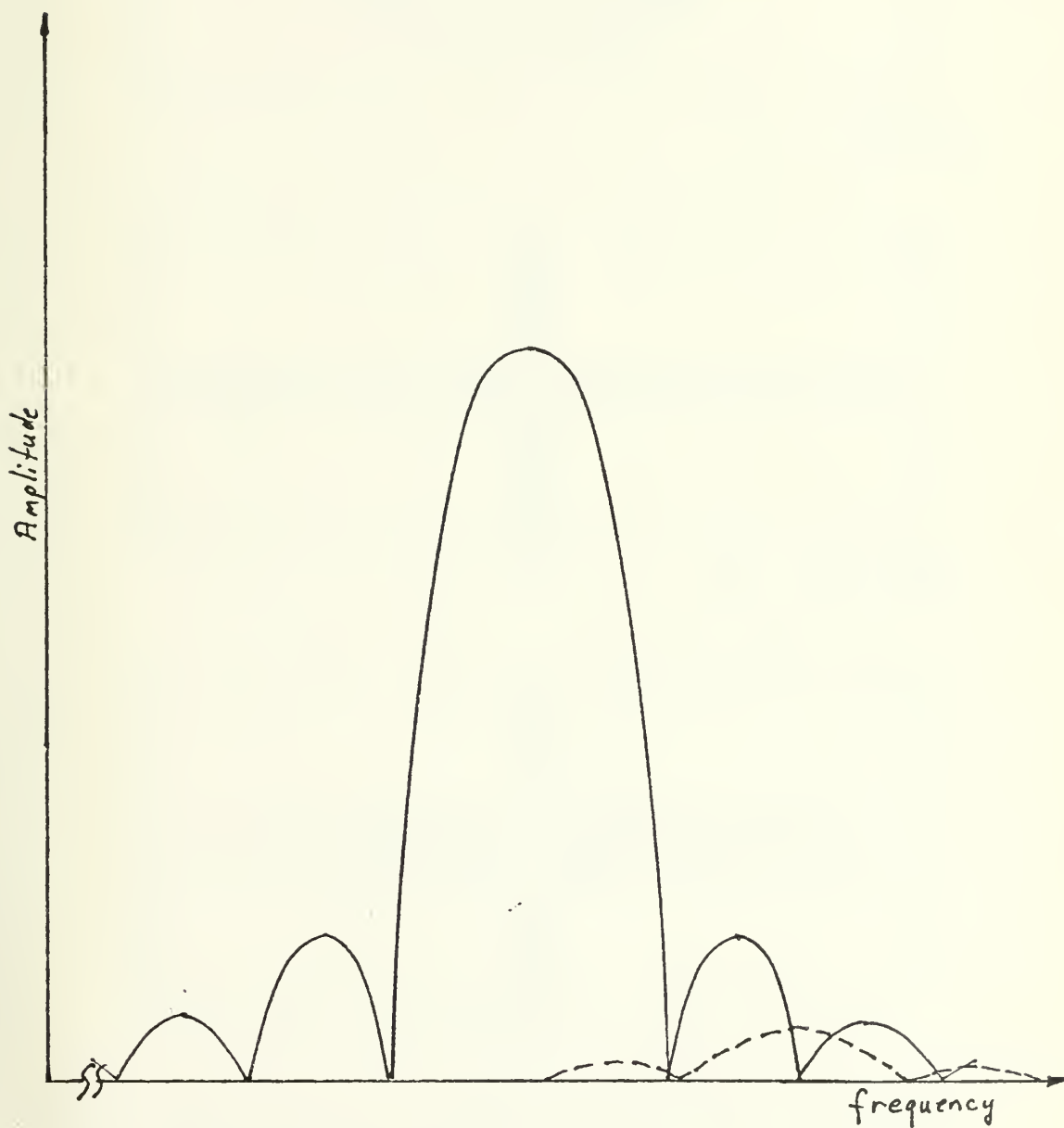
Figure 6   Effect of range sidelobes on weak signal detection

(a)

time

(b)

|← 5 μsec

(c)
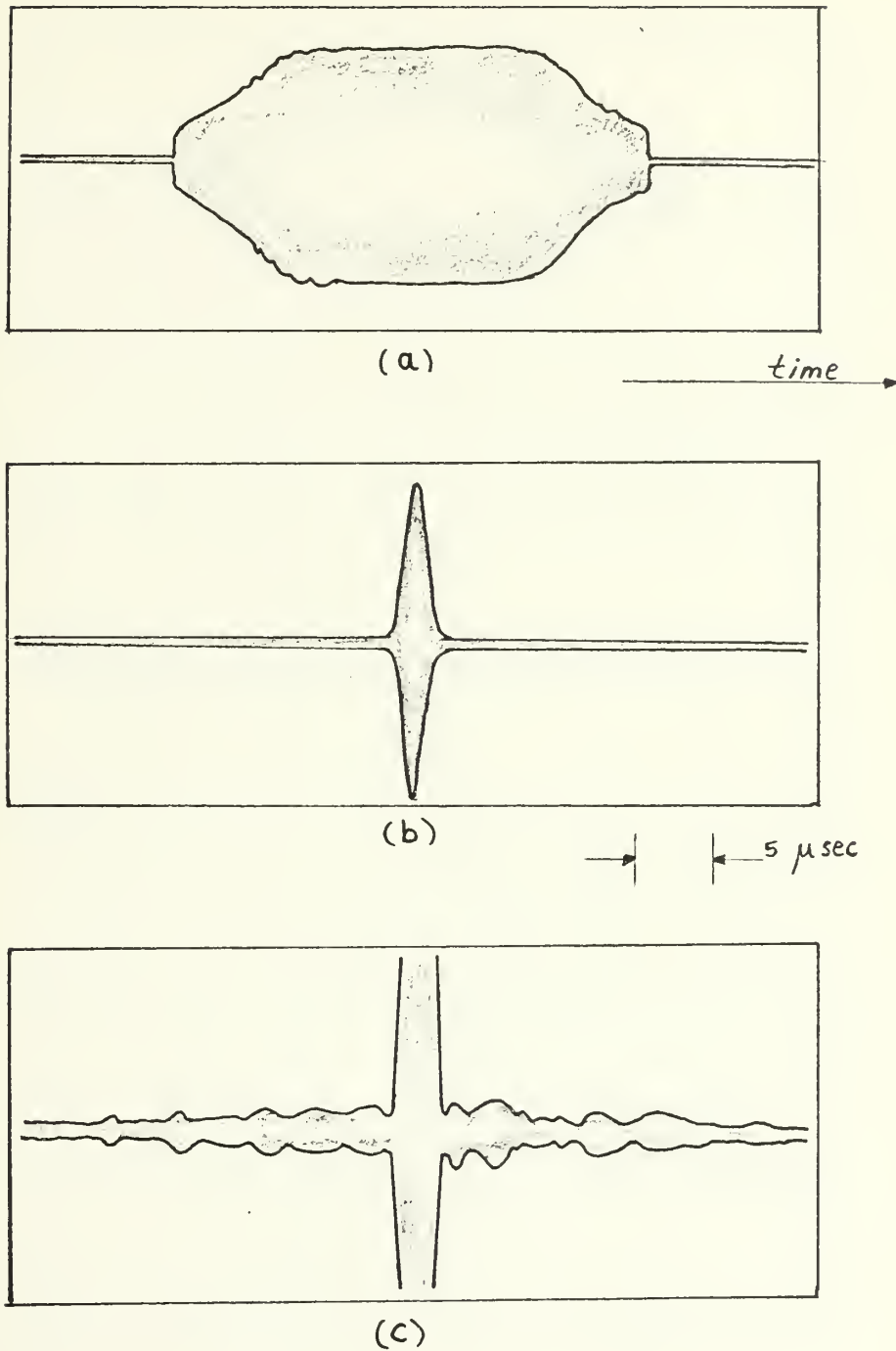
Figure 7   The receiver output waveform the truncated
Butterworth case

(a) Butterworth signal
(b) Receiver output waveform
(c) Receiver output waveform amplified 20 times

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(a)

time

(b)

5 μsec

(c)
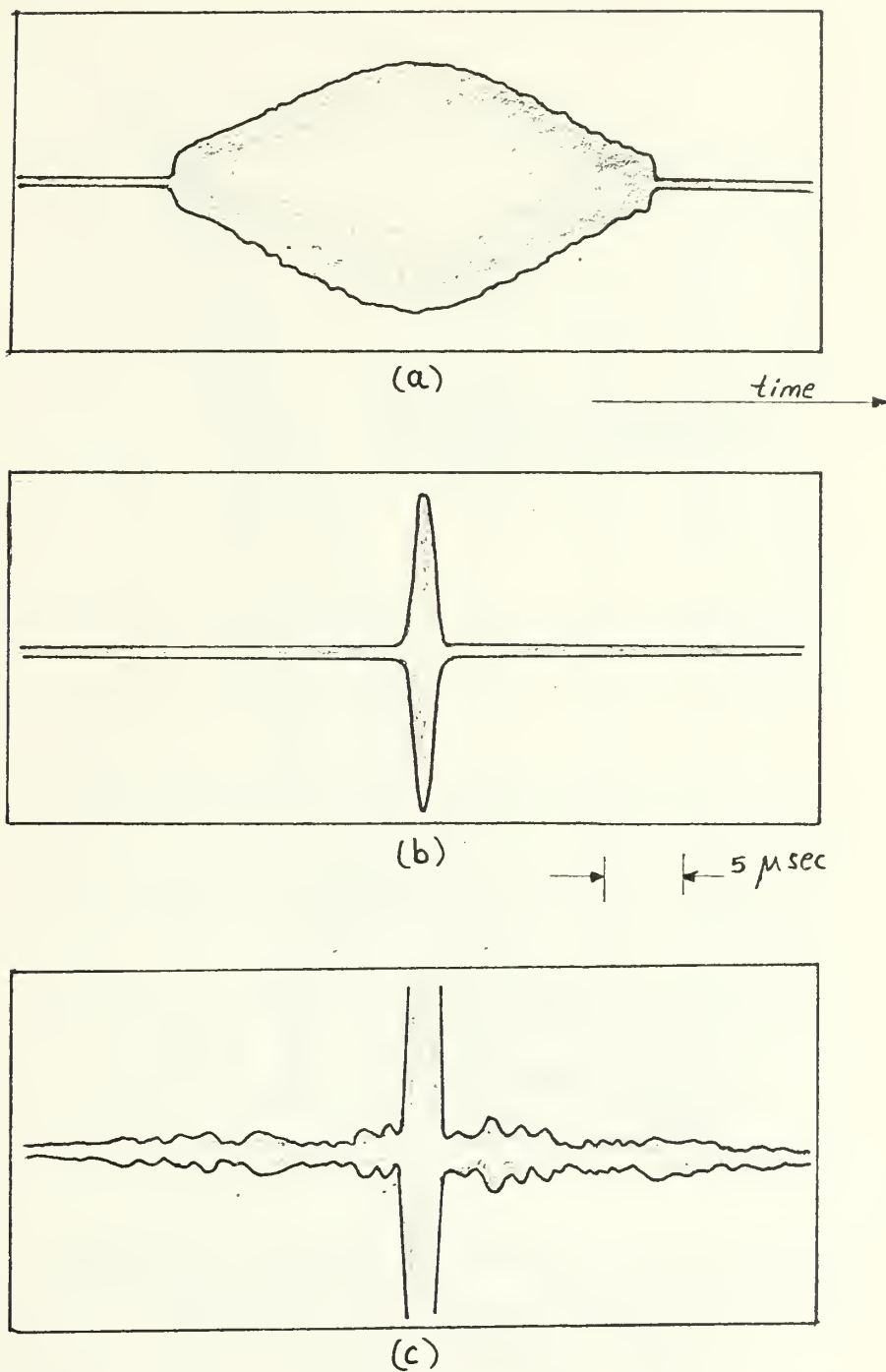
Figure 8  The receiver output waveform for the truncated
Gaussian case

(a) Gaussian signal

(b) Receiver output waveform

(c) Receiver output waveform amplified 20 times

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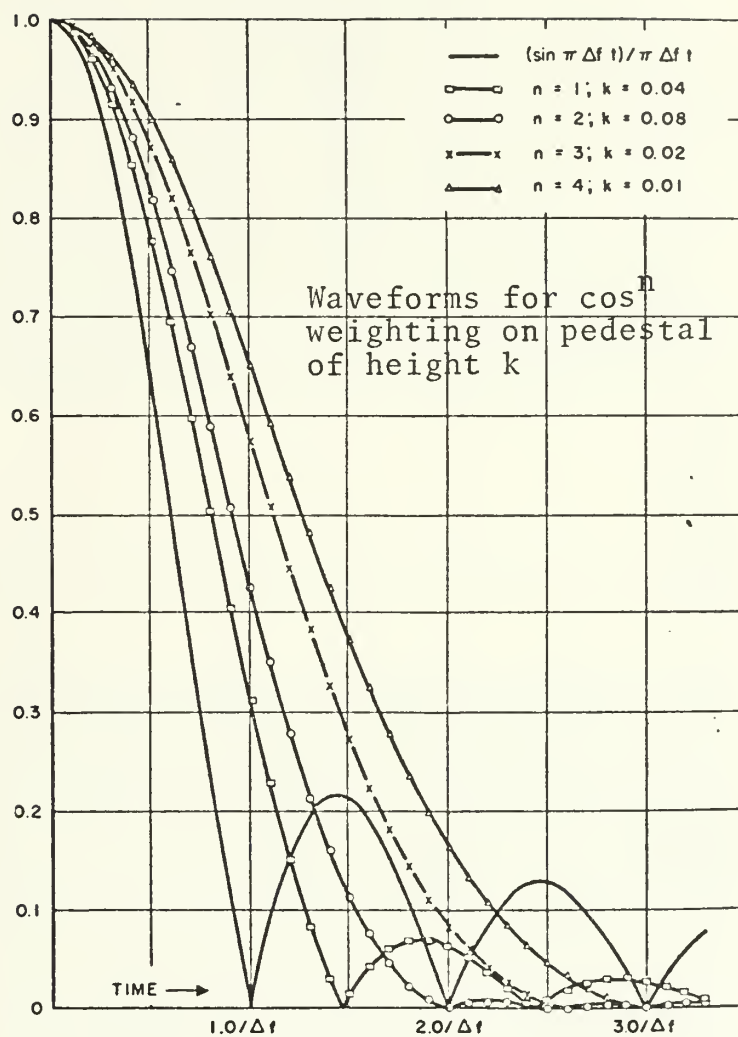
Figure 9   Typical waveforms resulting from
frequency weighting techniques
at the matched-filter output
[Refs. 2, 3]

30
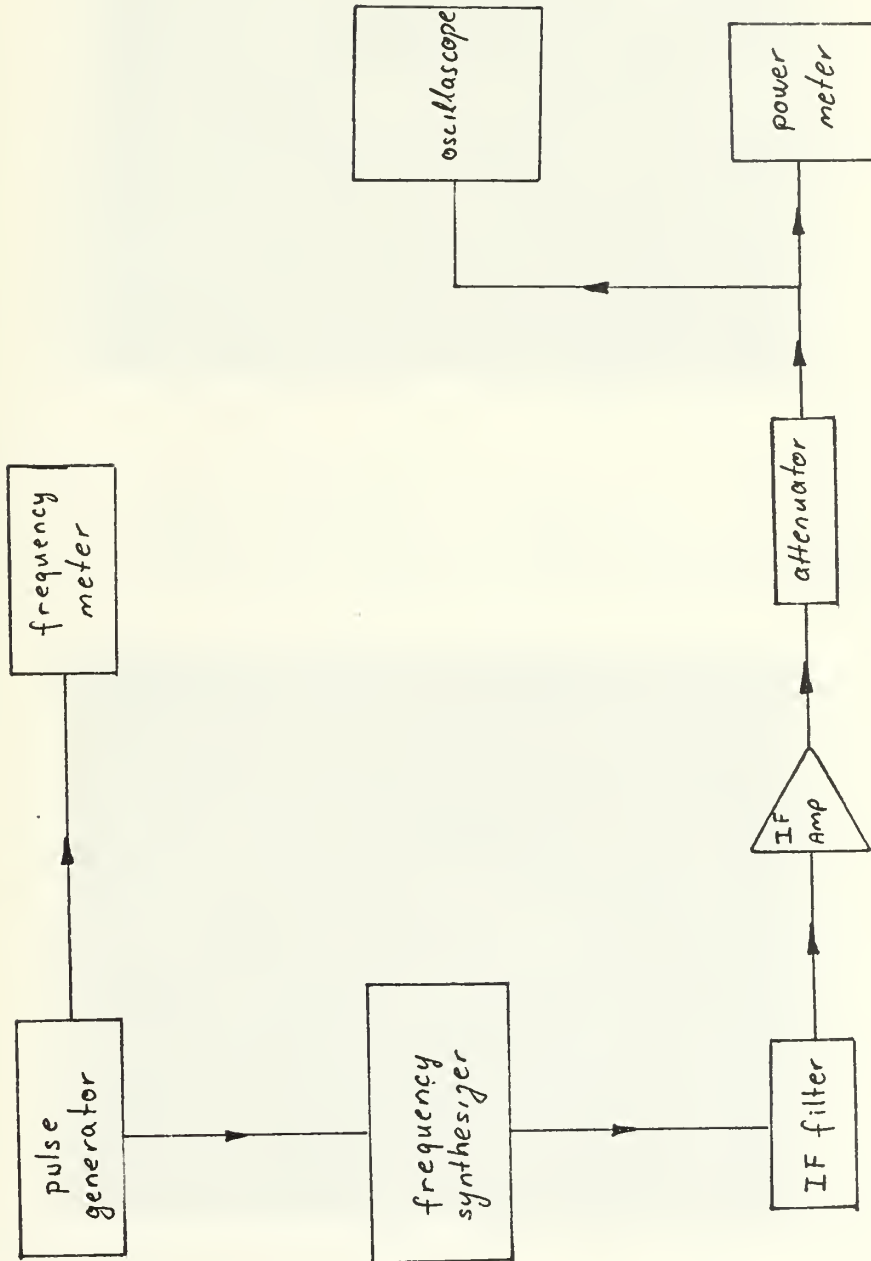
Figure 10 Block diagram of the experimental setup
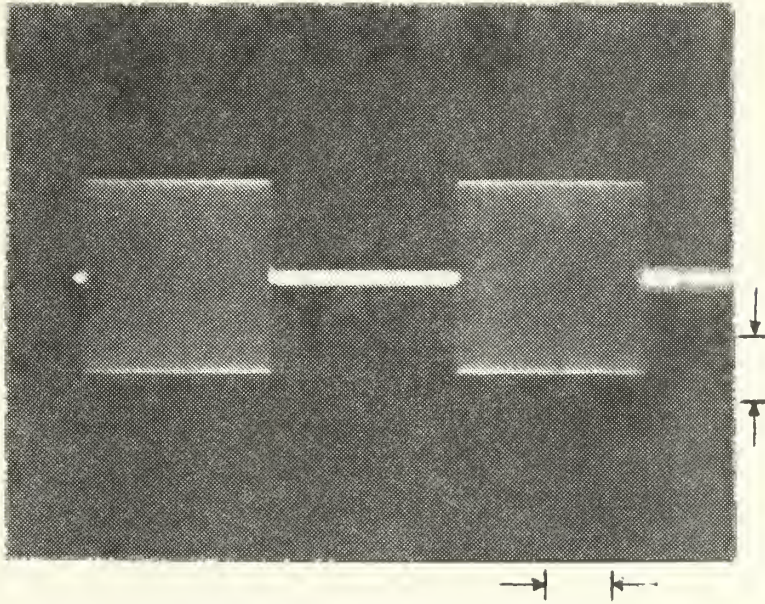
Figure 11  Pulse modulated IF carrier
        Vertical scale  :  1 volt per division
        Horizontal scale:  100 μsec per division
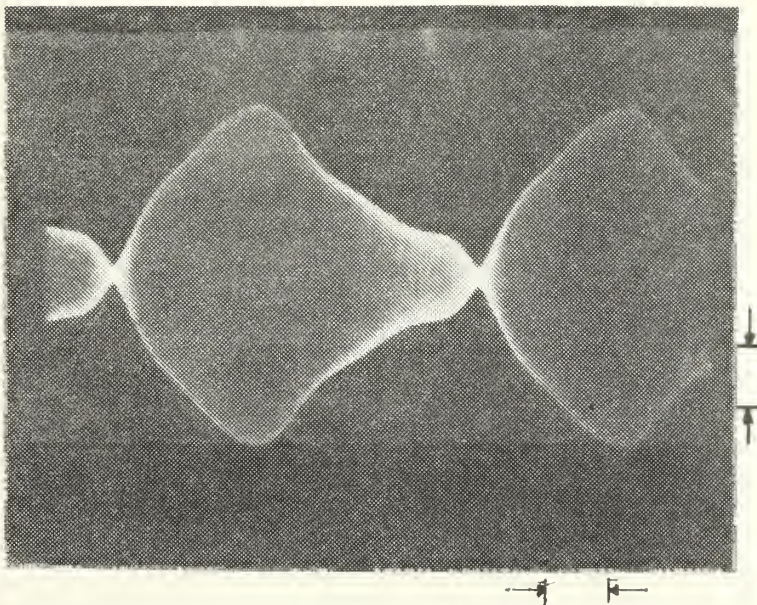


Figure 12  IF filter response
        Vertical scale  :  0,01 volts per division
        Horizontal scale:  200 μsec per division
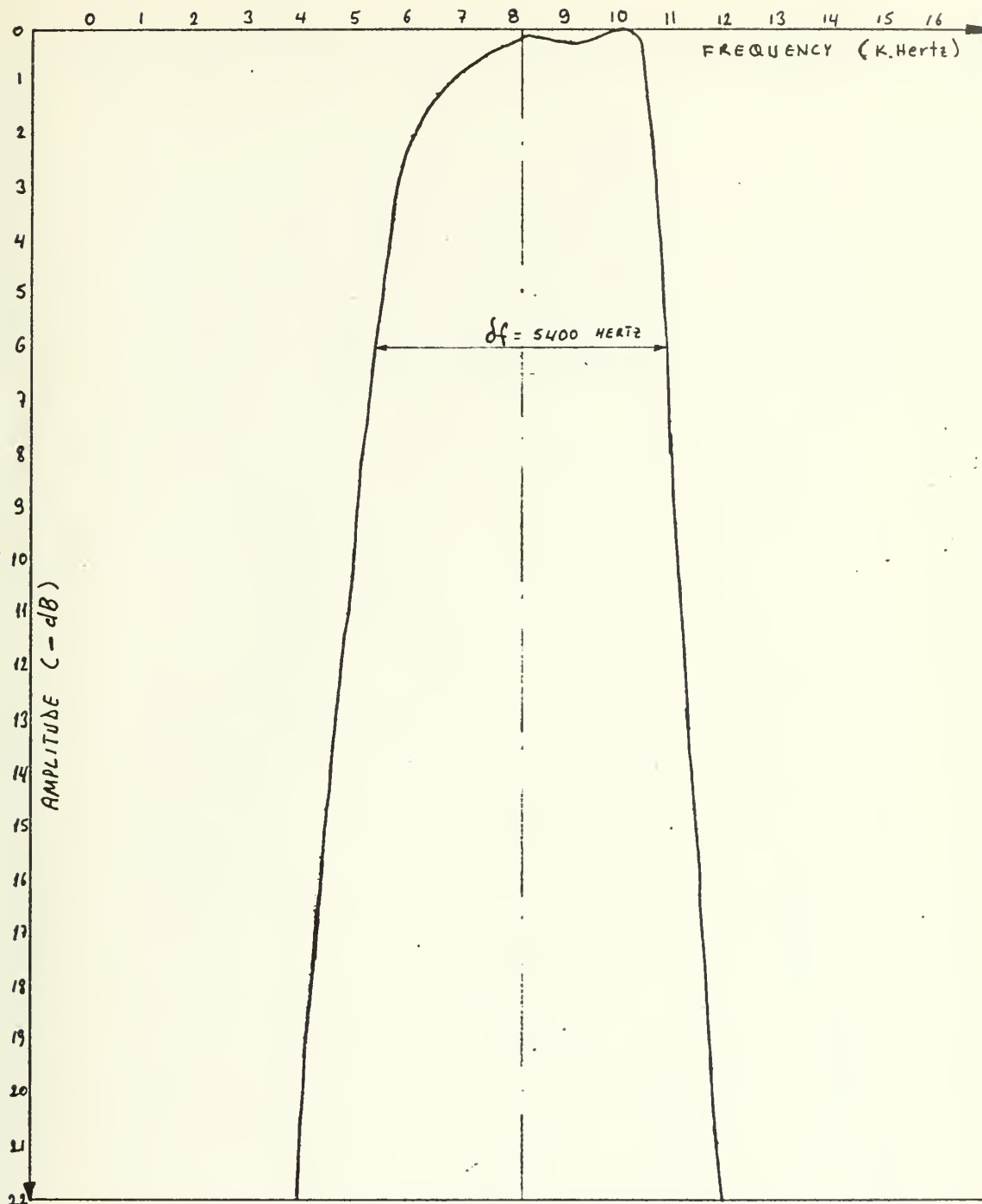
Figure 13    High-Q quartz crystal narrow bandpass IF filter
response

Center frequency:   29,998,200 Hertz

δf:                 5400 Hertz

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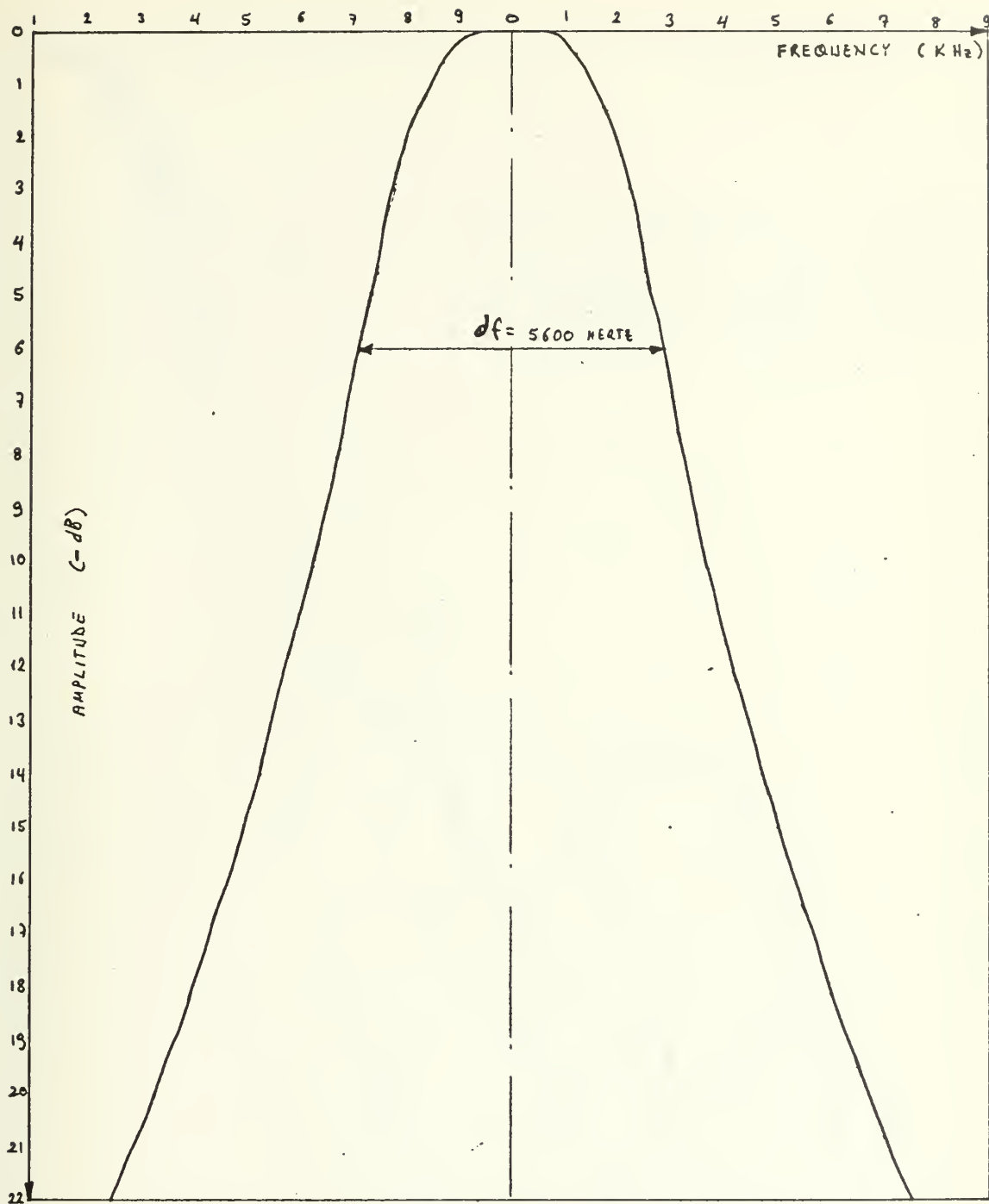
Figure 14   Analytical IF filter characteristic

Center frequency:   30,000,000 Hertz

δf:                              5600 Hertz

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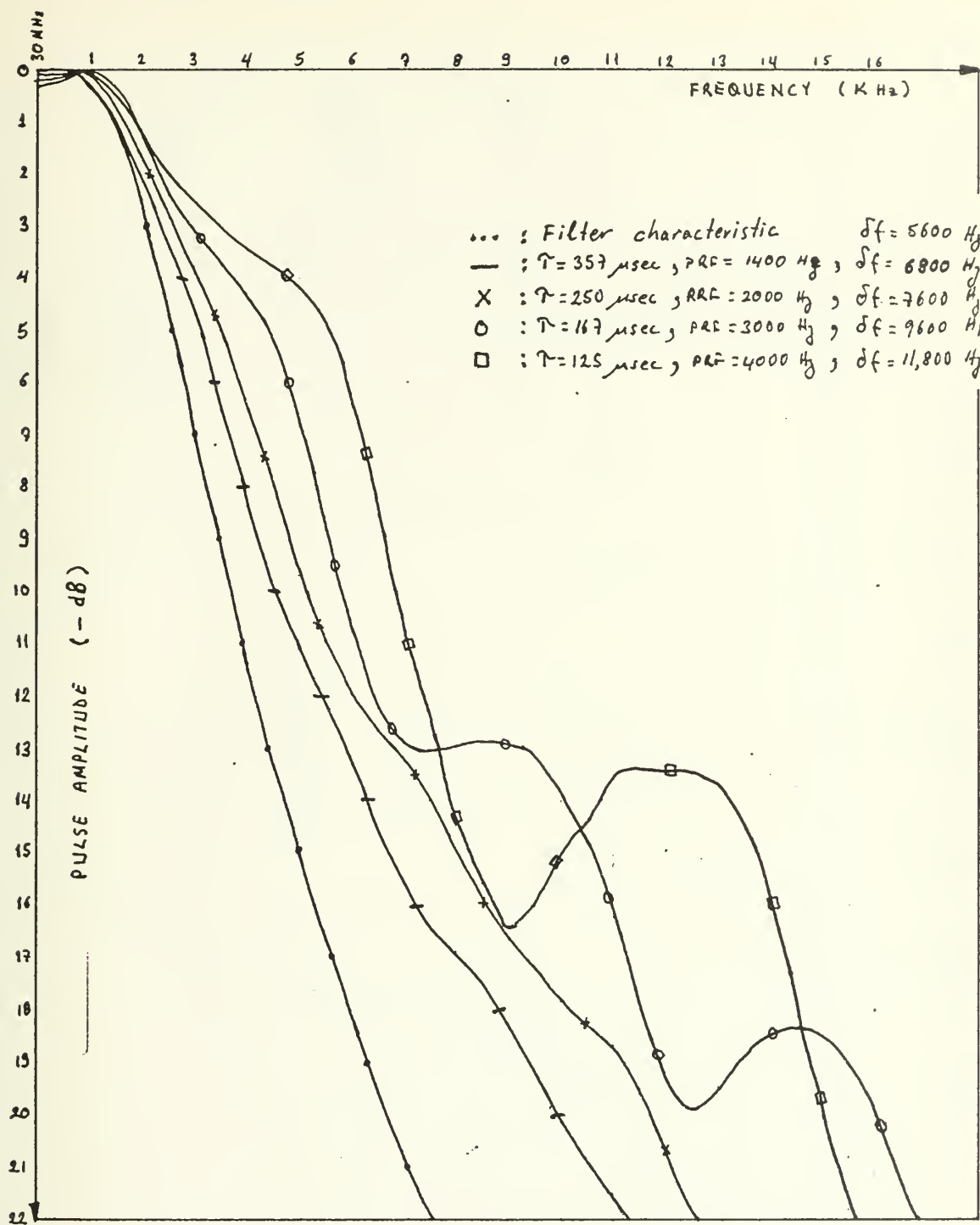
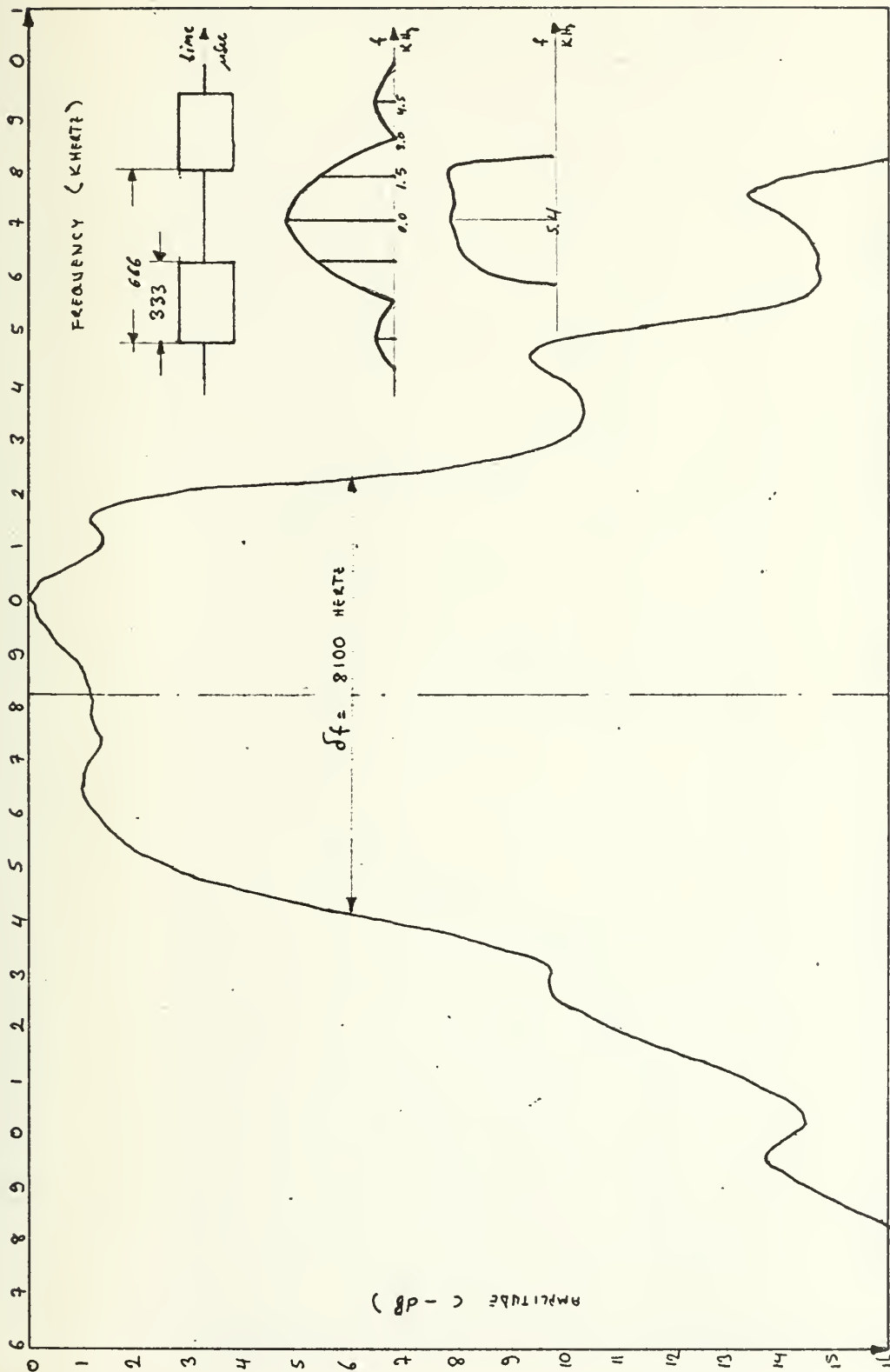Figure 15  Effects of the pulsewidth of the pulse signal
on the IF filter
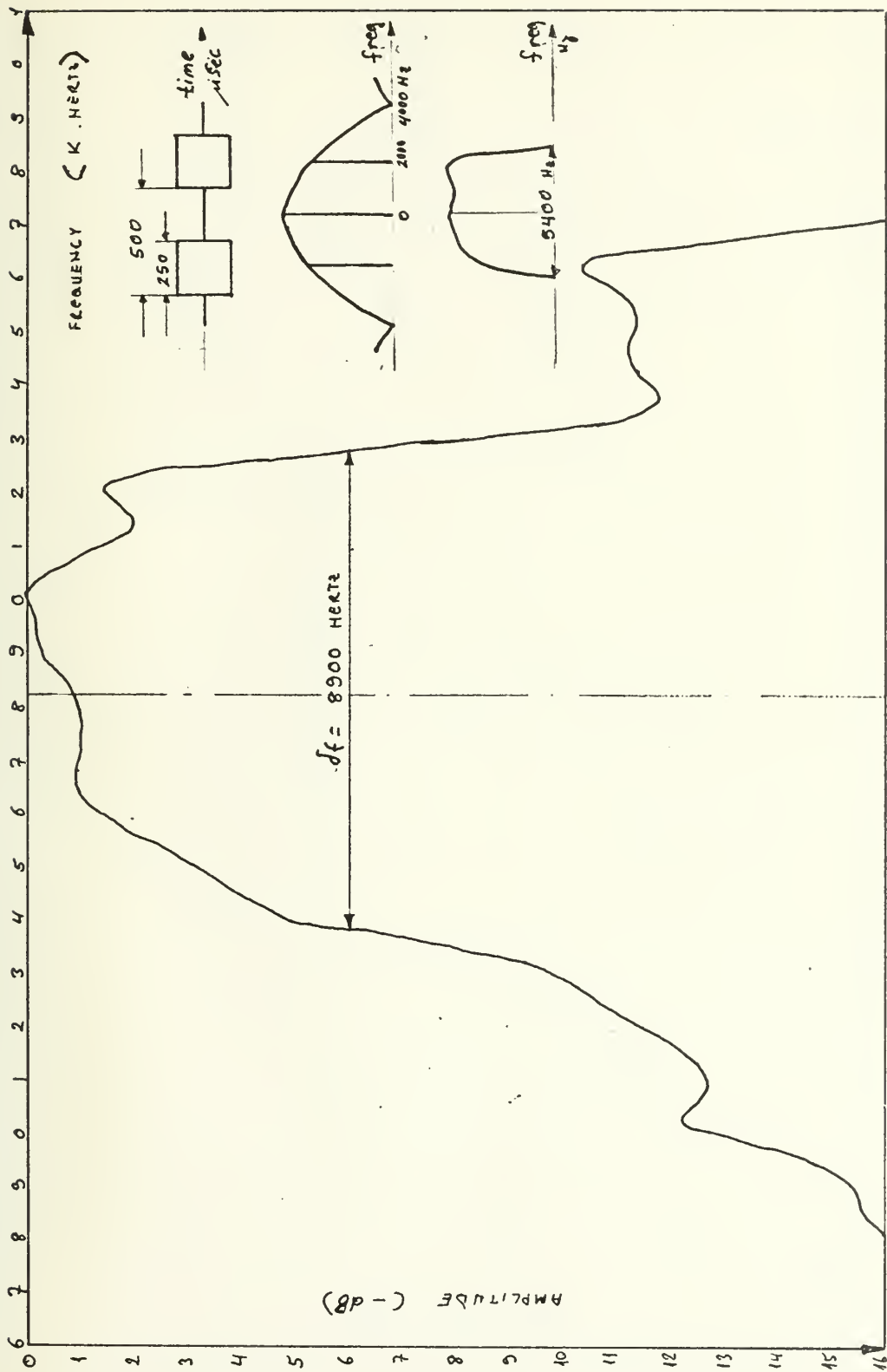
Figure 16 The effect of the width of the input pulse signal on the IF filter
First sidelobe 10.3 dB lower than the mainlobe
Input pulse width τ = 250 μsec
Duty cycle = 1/2

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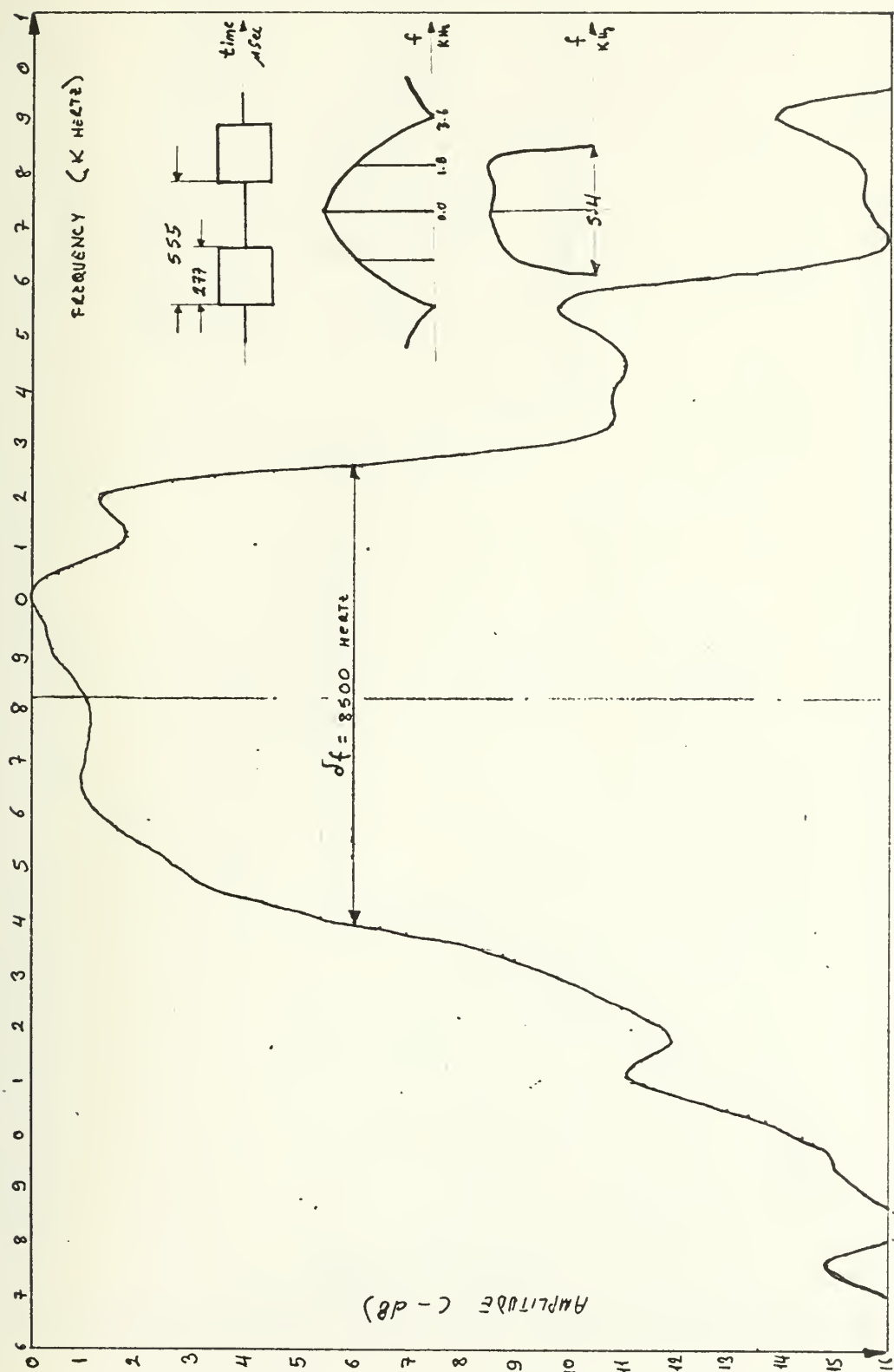
Figure 17 The effect of the width of the input pulse signal on the IF filter
First sidelobe 9.8 dB lower than the mainlobe
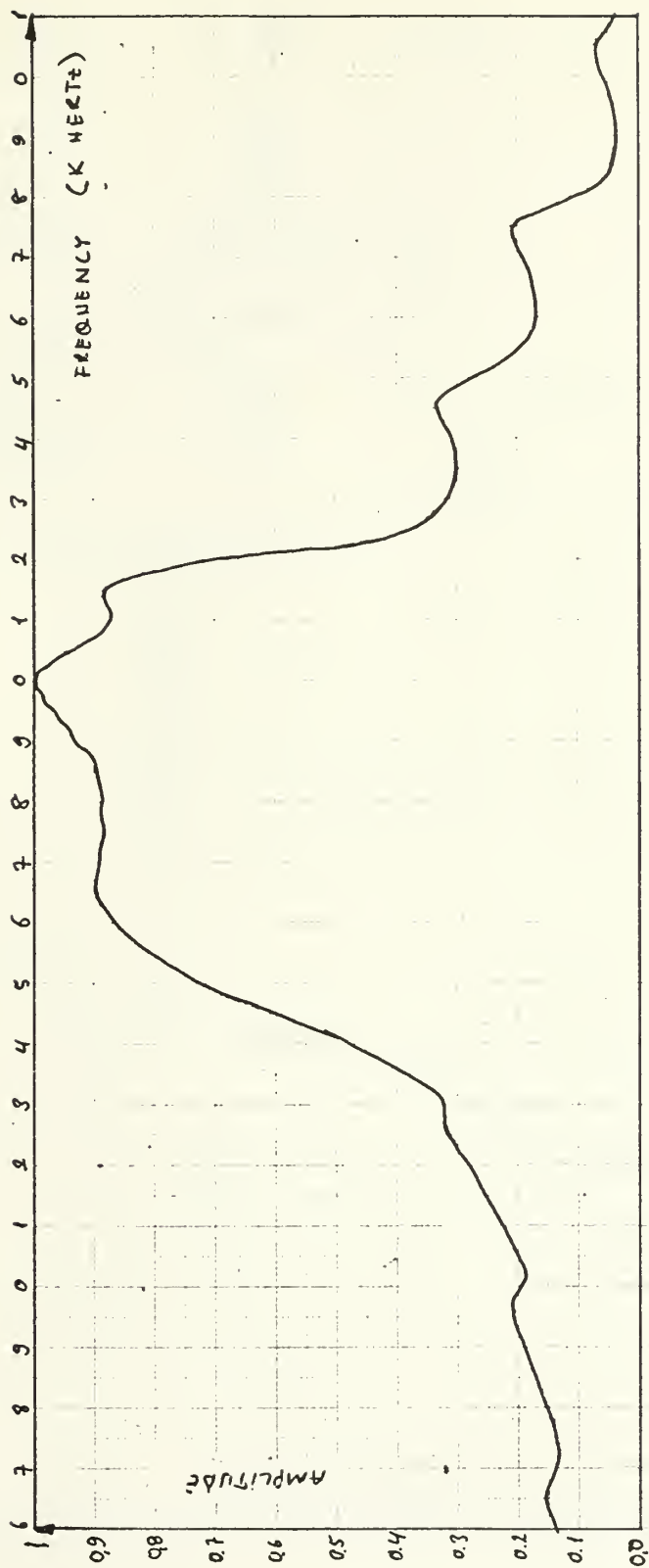Input pulse width $\tau$ = 277 µsec
Duty cycle = 1/2

37
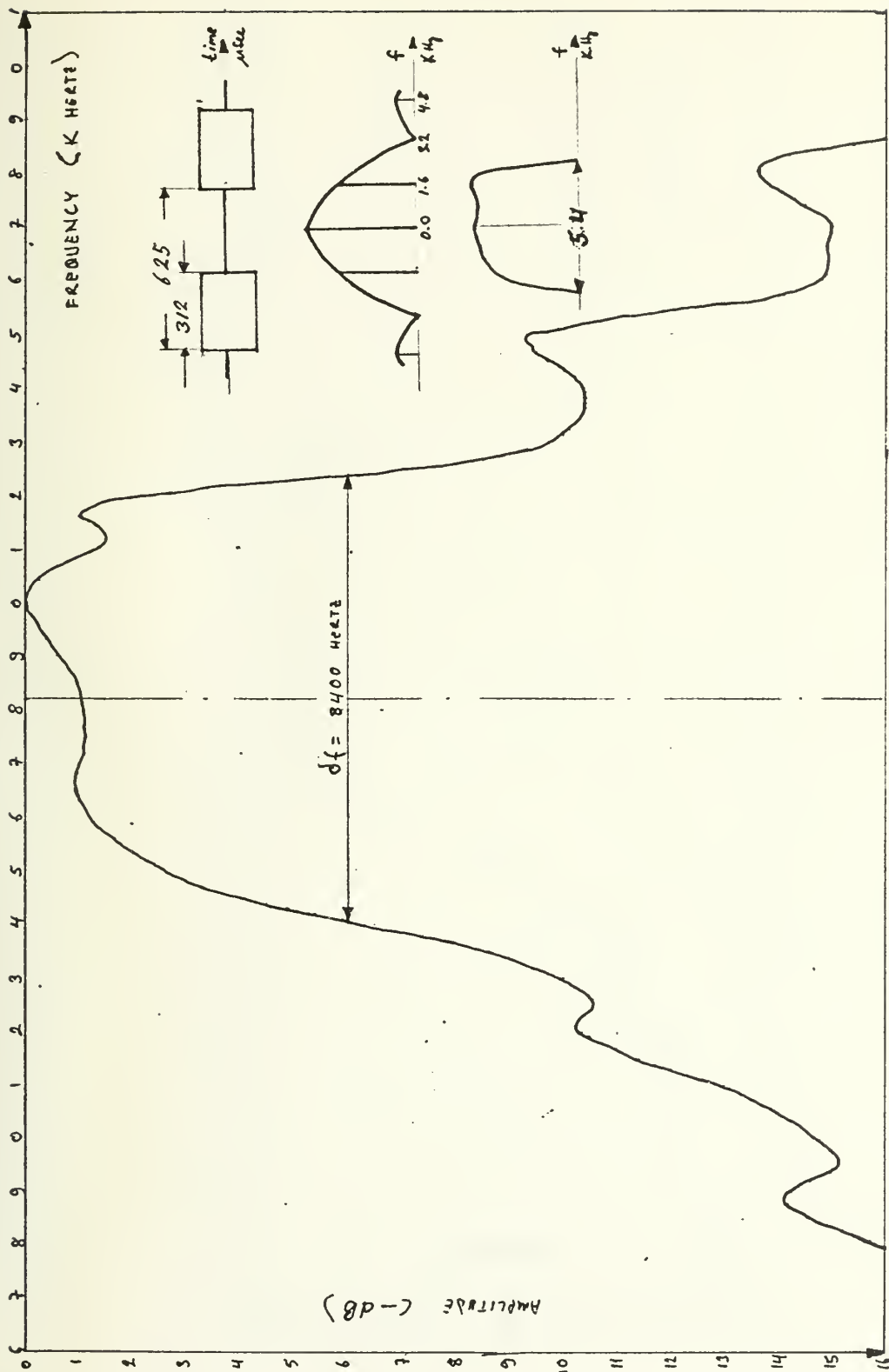
Figure 18 (a)  The effect of the widt- of the input pulse signal
on the IF filter
First sidelobe 9.3 dB lower than the mainlobe
Input pulse width τ = 312 μsec
Duty cycle = 1/2

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Figure 18 (b)   The effect of the width of the input pulse signal on the IF filter
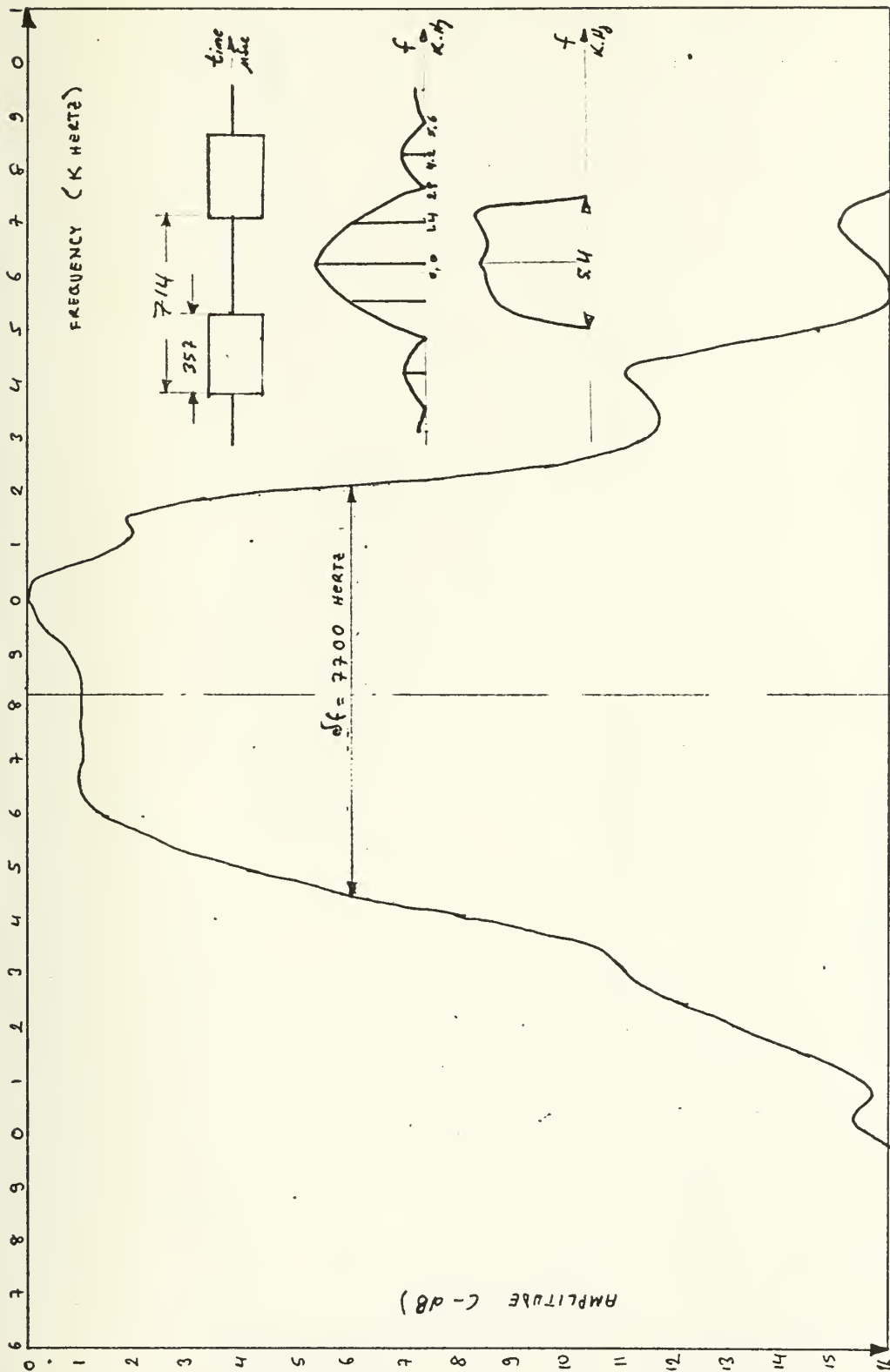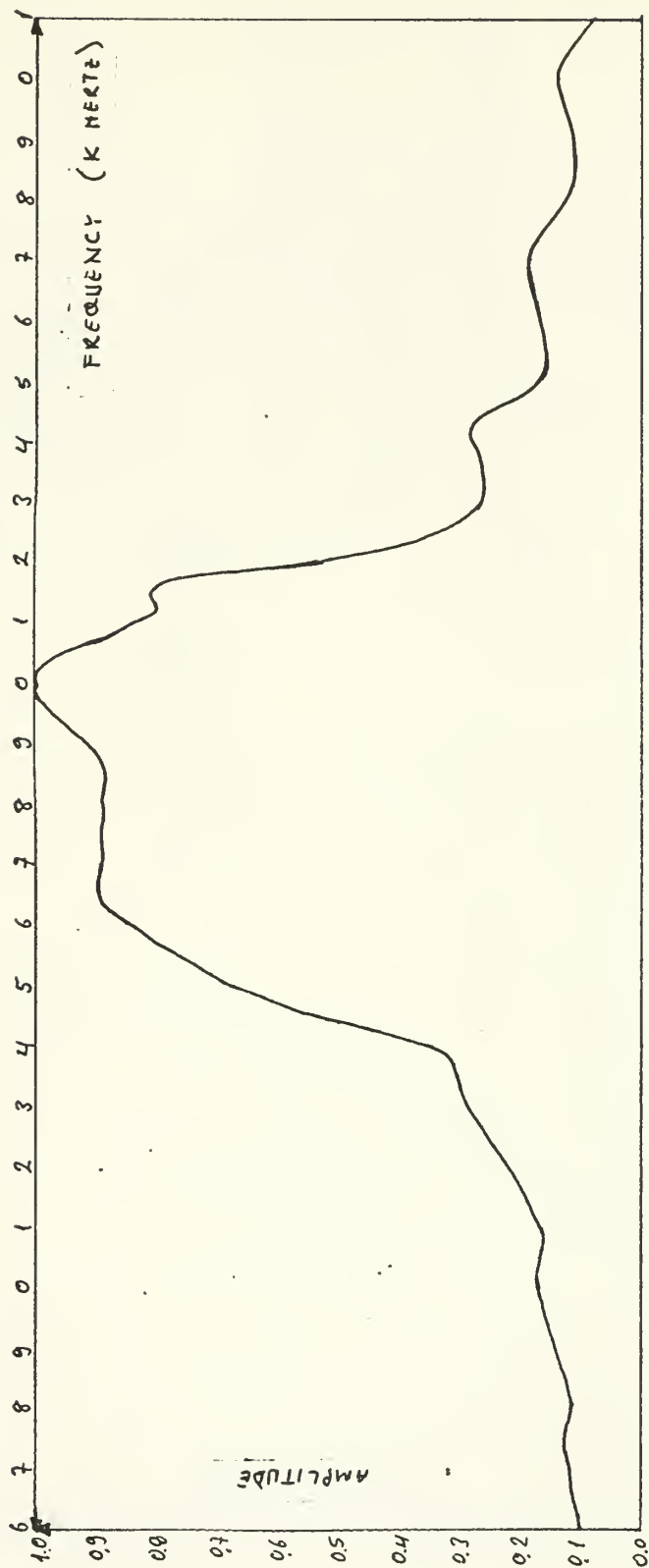Figure 18(b) shows the same information on a linear scale for
comparison.

39

Figure 19 (a)   The effect of the width of the input pulse signal on
the IF filter
First sidelobe 9.3 dB lower than the mainlobe
Input pulse width $\tau$ = 333 $\mu$sec
Duty cycle = 1/2

40

Figure 19 (b)   The effect of the width of the input pulse signal on
the IF filter
Figure 19(b) shows the same information on a linear
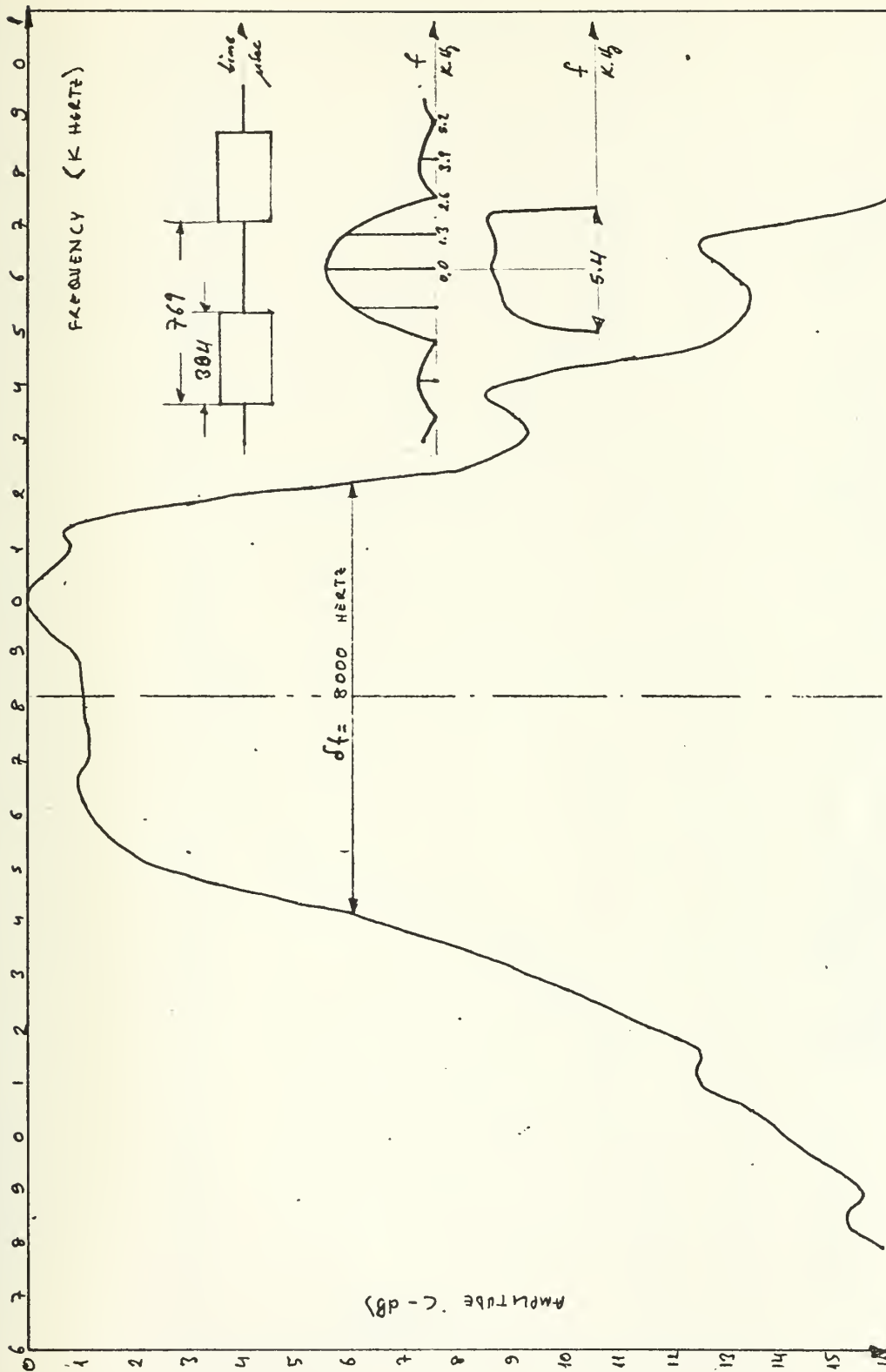scale for comparison.

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Figure 20 (a)   The effect of the width of the input pulse signal on
the IF filter
First sidelobe 11.1 dB lower than the mainlobe
Input pulse width $\tau$ = 357 µsec
Duty cycle = 1/2

Figure 20 (b)    The effect of the width of the input pulse signal
                 on the IF filter
                 Figure 20(b) shows the same information on a linear
                 scale for comparison.

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Figure 21 (a)   The effect of the width of the input pulse signal on
the IF filter
First sidelobe 8.4 dB lower than the mainlobe
Input pulse width $\tau$ = 384 μsec
Duty cycle = 1/2

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Figure 21 (b)  The effect of the width of the input pulse signal on the IF filter Figure 21(b) shows the same information on a linear scale for comparison.
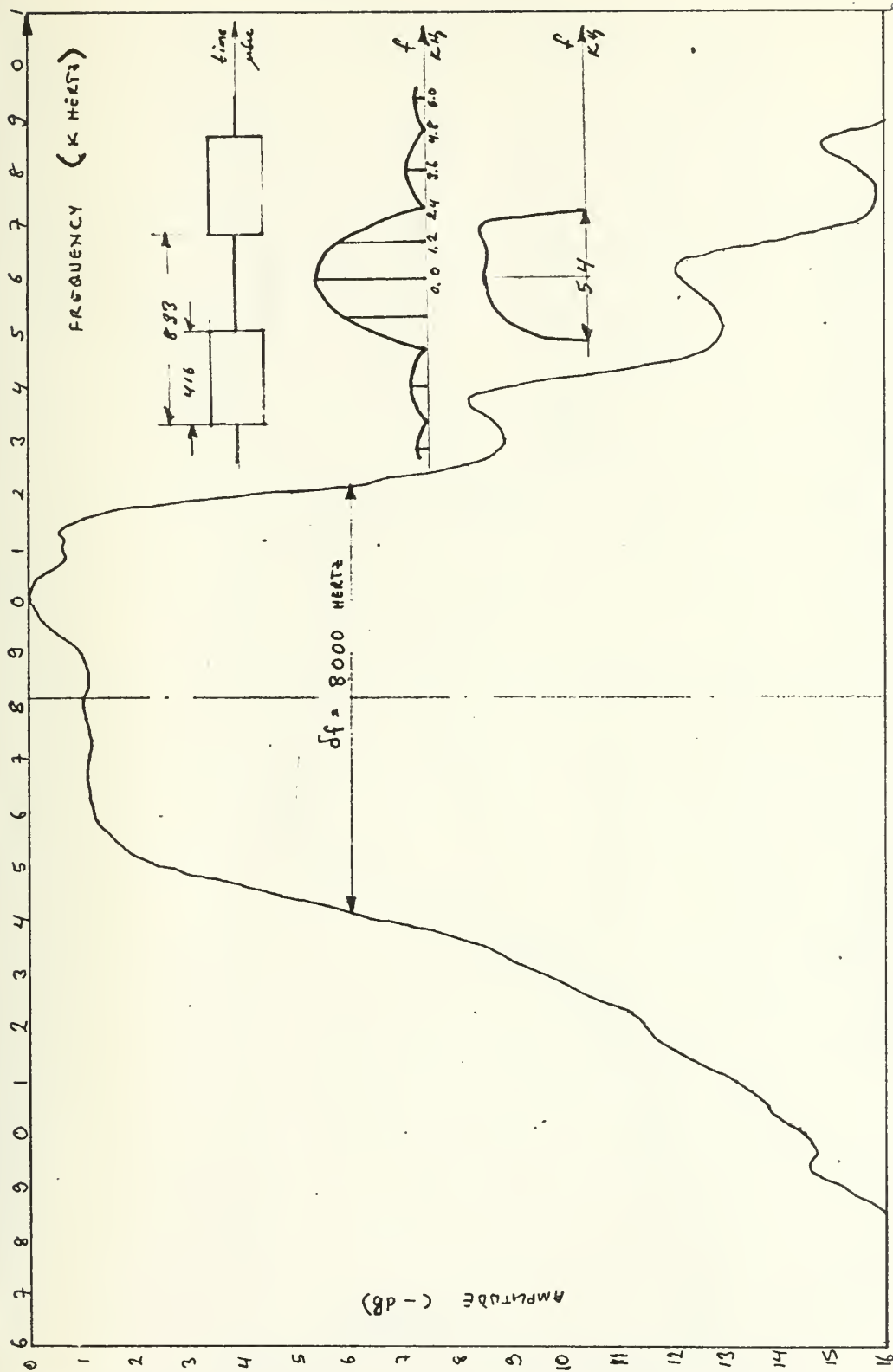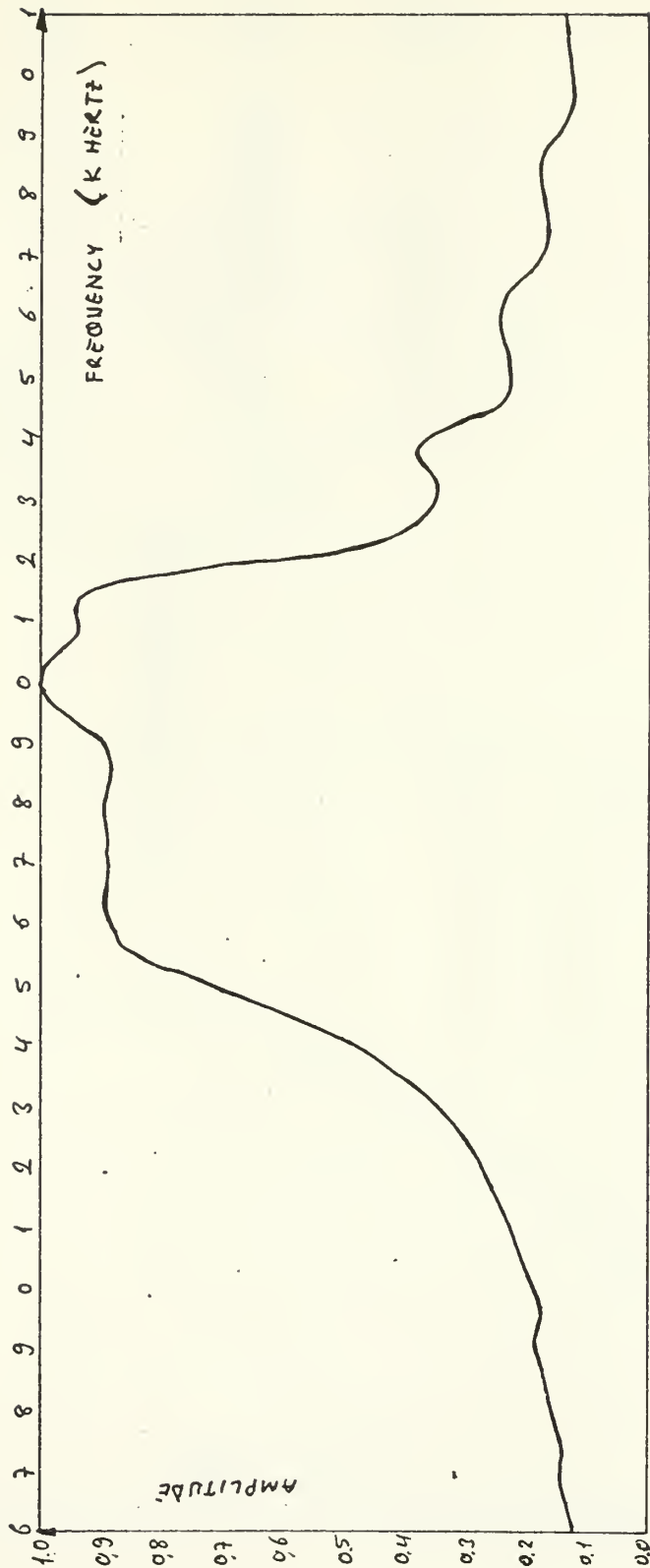
Figure 22 (a)   The effect of the width of the input pulse signal on
                the IF filter
                First sidelobe 8.2 dB lower than the mainlobe
                Input pulse width τ = 416 μsec
                Duty cycle = 1/2

Figure 22 (b)   The effect of the width of the input pulse signal on the IF filter Figure 22(b) shows the same information on a linear scale for comparison.
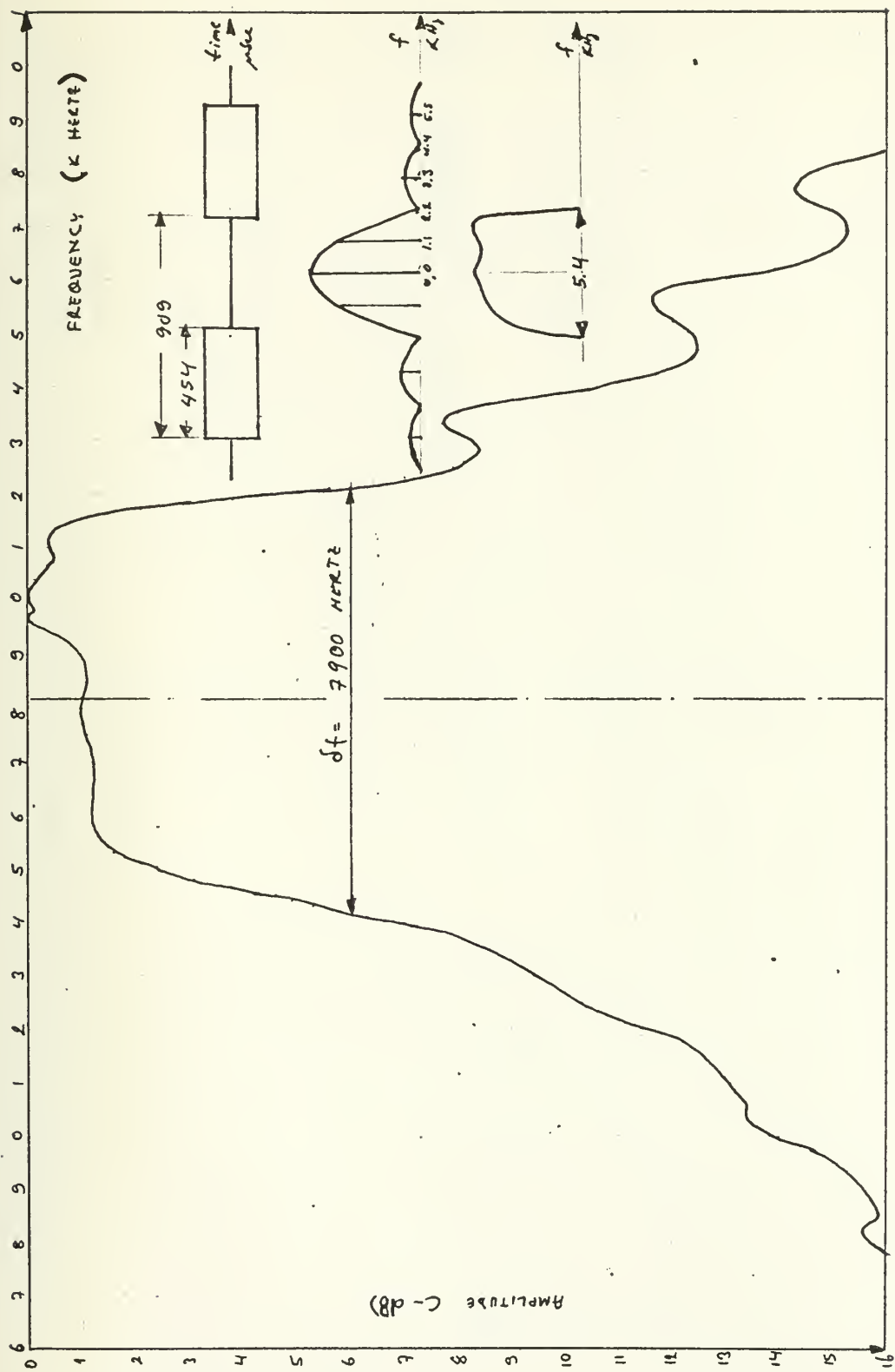
Figure 23  The effect of the width of the input pulse signal on the
IF filter
First sidelobe 7.7 dB lower than the mainlobe
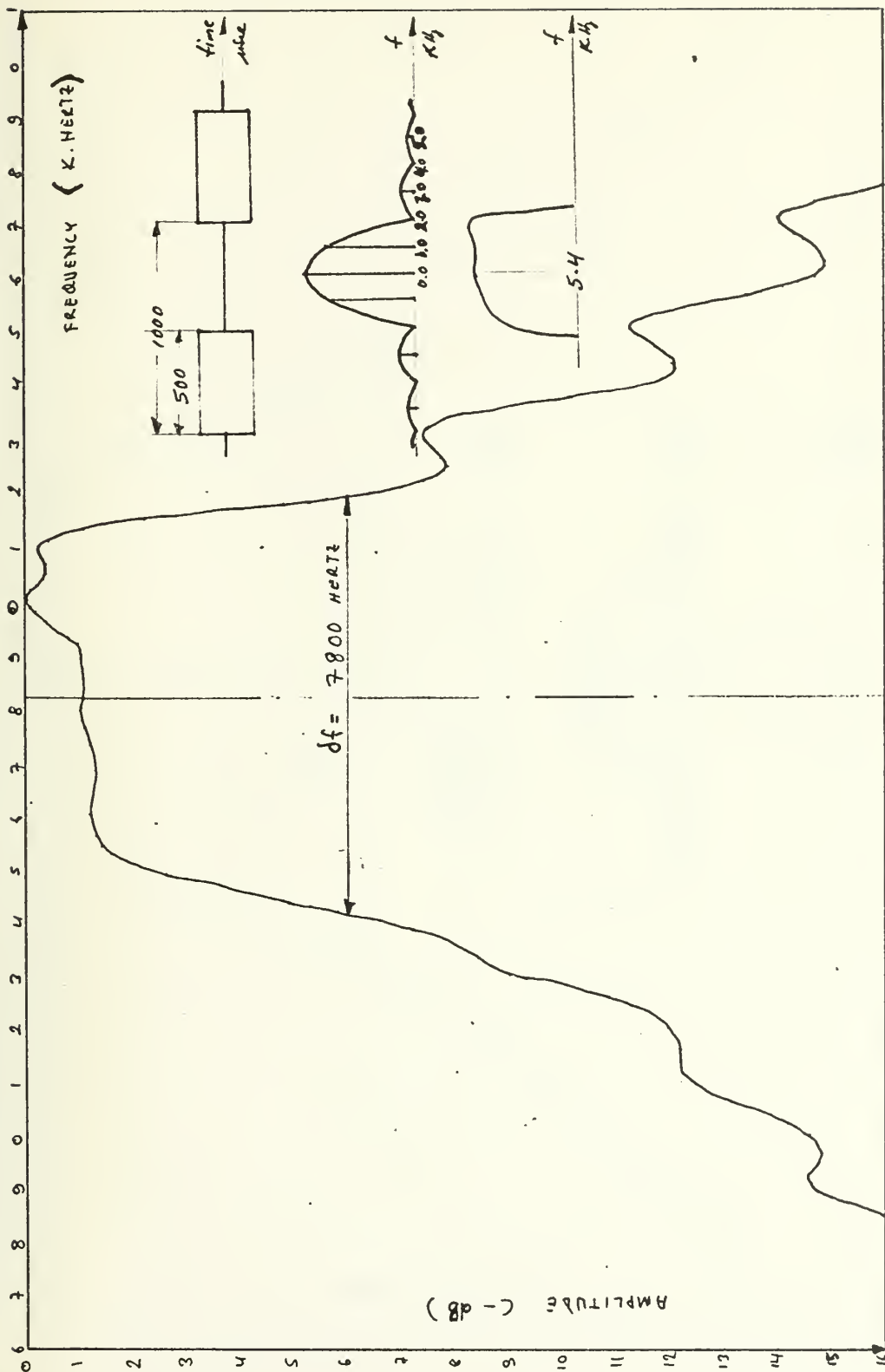Input pulse width τ = 454 μsec
Duty cycle = 1/2

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Figure 24 The effect of the width of the input pulse signal on the
IF filter
First sidelobe 7.4 dB lower than the mainlobe
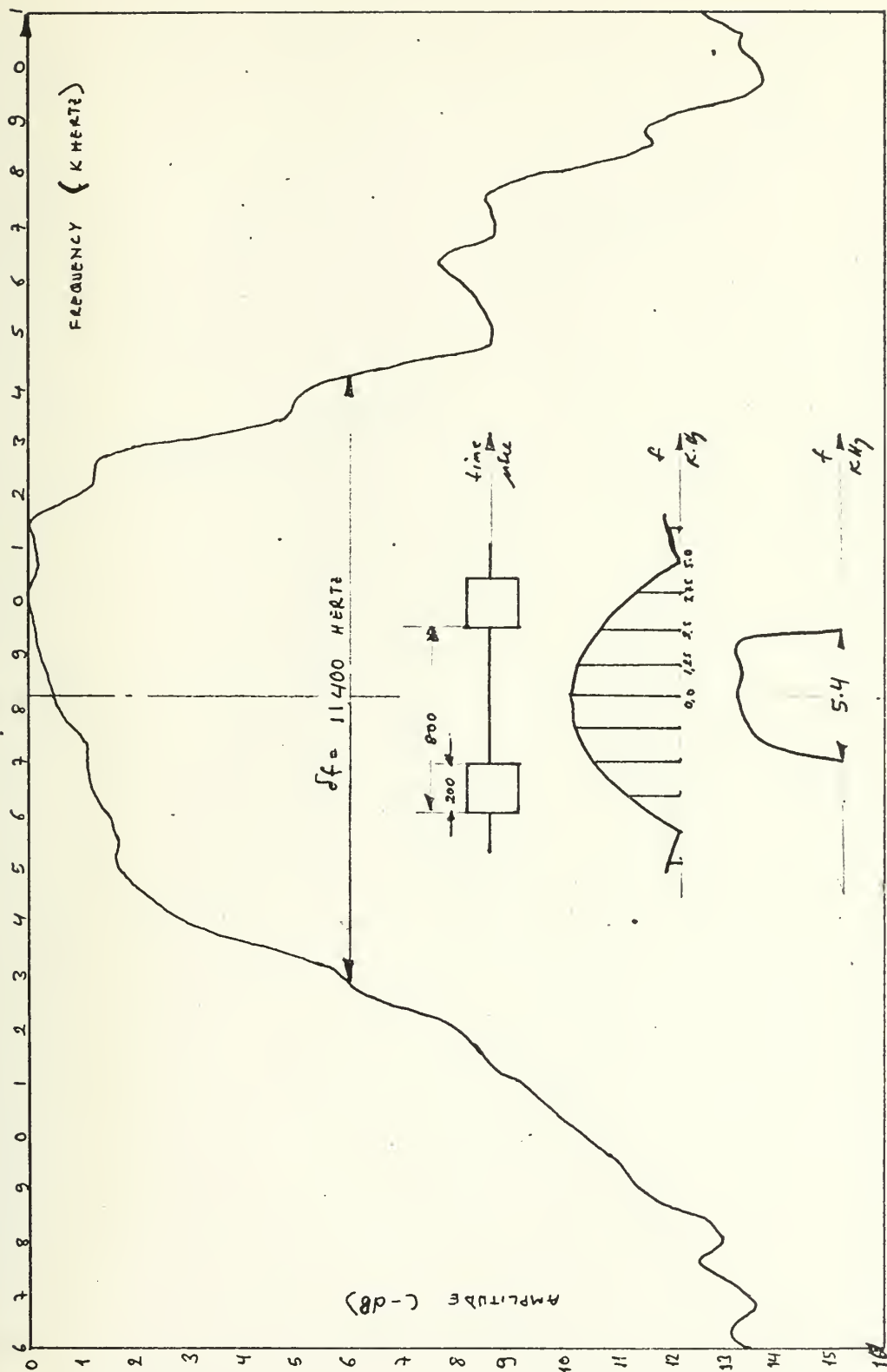Input pulse width τ = 500 μsec
Duty cycle = 1/2

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Figure 25 The effect of the width of the input pulse signal on the
IF filter
First sidelobe 7.7 dB lower than the mainlobe
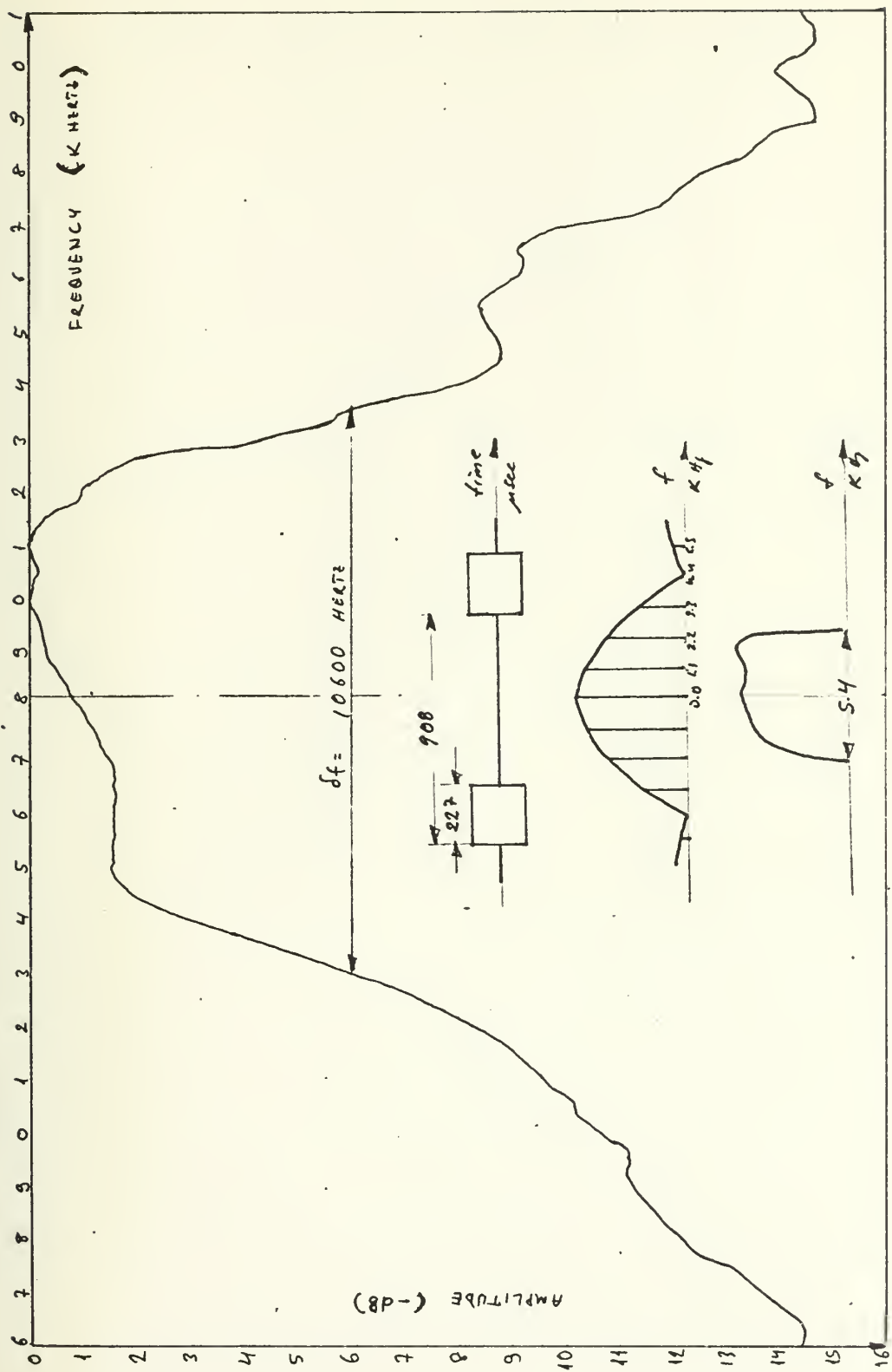Input pulse width τ = 200 μsec
Duty cycle = 1/4

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Figure 26   The effect of the width of the input pulse signal on the
IF filter
First sidelobe 8.4 dB lower than the mainlobe
Input pulse width τ = 227 μsec
Duty cycle = 1/4

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Figure 27 The effect of the width of the input pulse signal on the
IF filter
First sidelobe 7.2 dB lower than the mainlobe
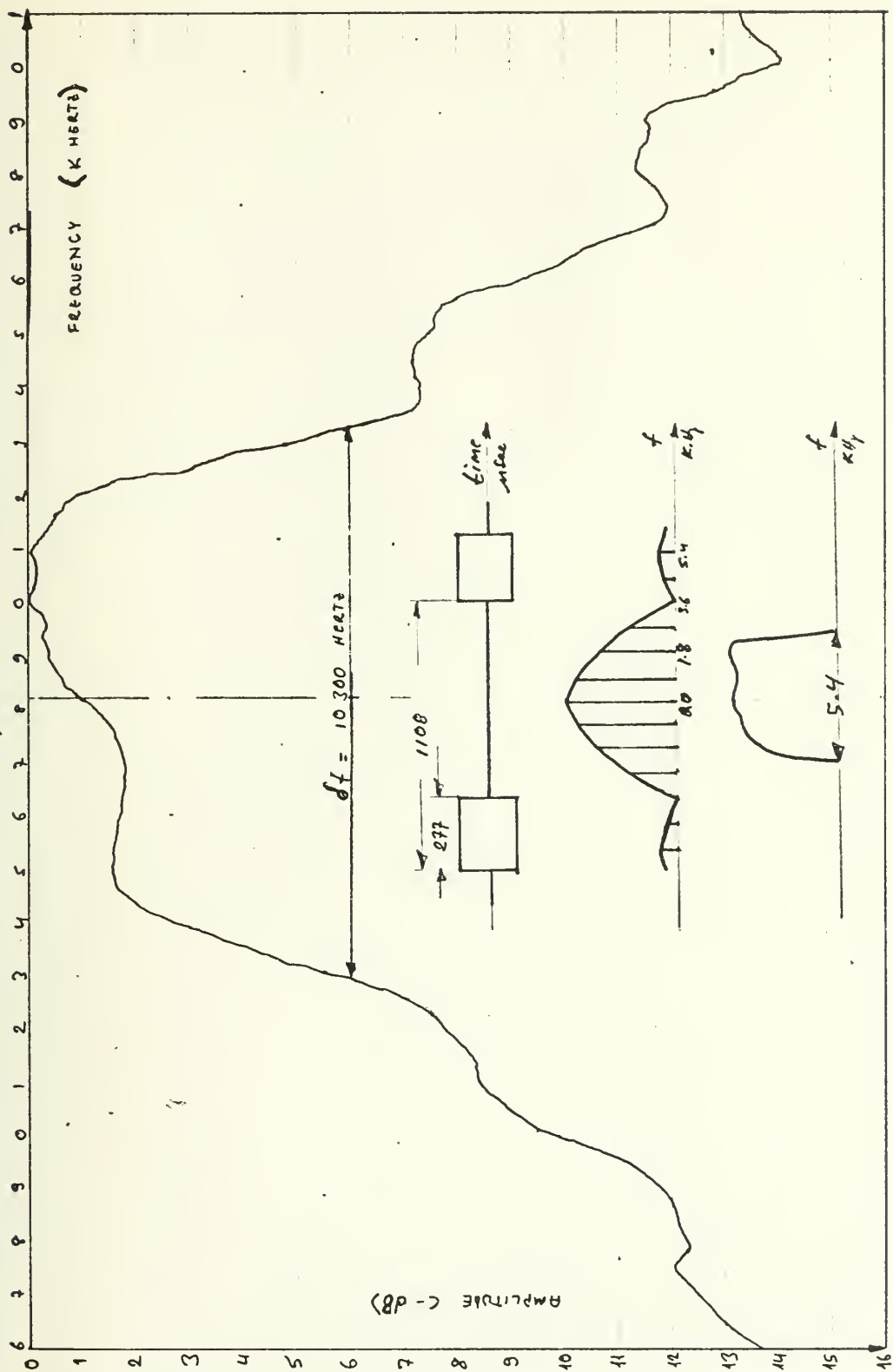Input pulse width τ = 277 μsec
Duty cycle = 1/4

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Figure 28    The effect of the width of the input pulse signal on the
IF filter
First sidelobe 7.3 dB lower than the mainlobe
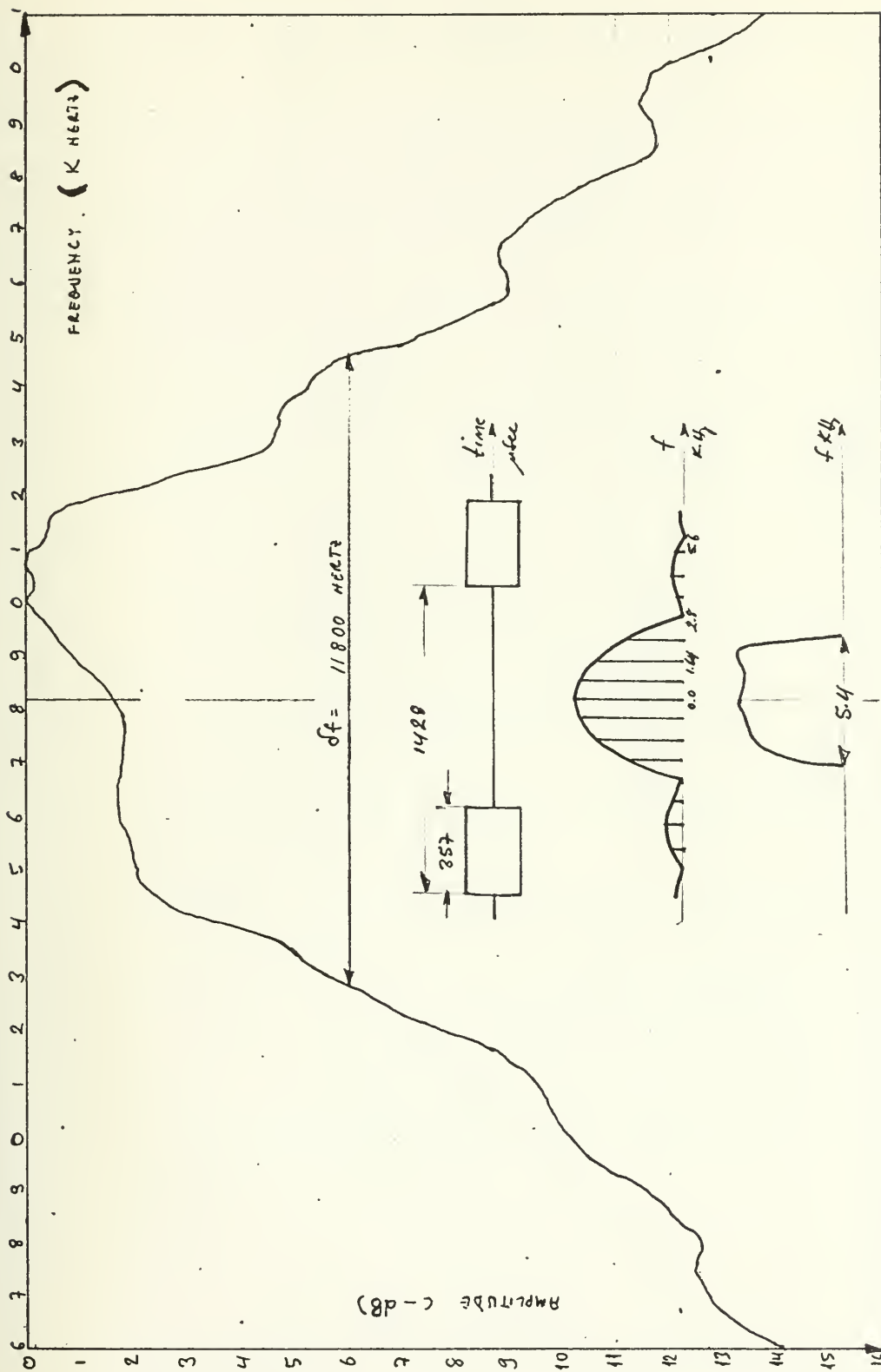Input pulse width $\tau$ = 312.5 µsec
Duty cycle = 1/4

Figure 29  The effect of the width of the input pulse signal on the
IF filter
First sidelobe 4.7 dB lower than the mainlobe
Input pulse width $\tau$ = 357 µsec
Duty cycle = 1/4

54
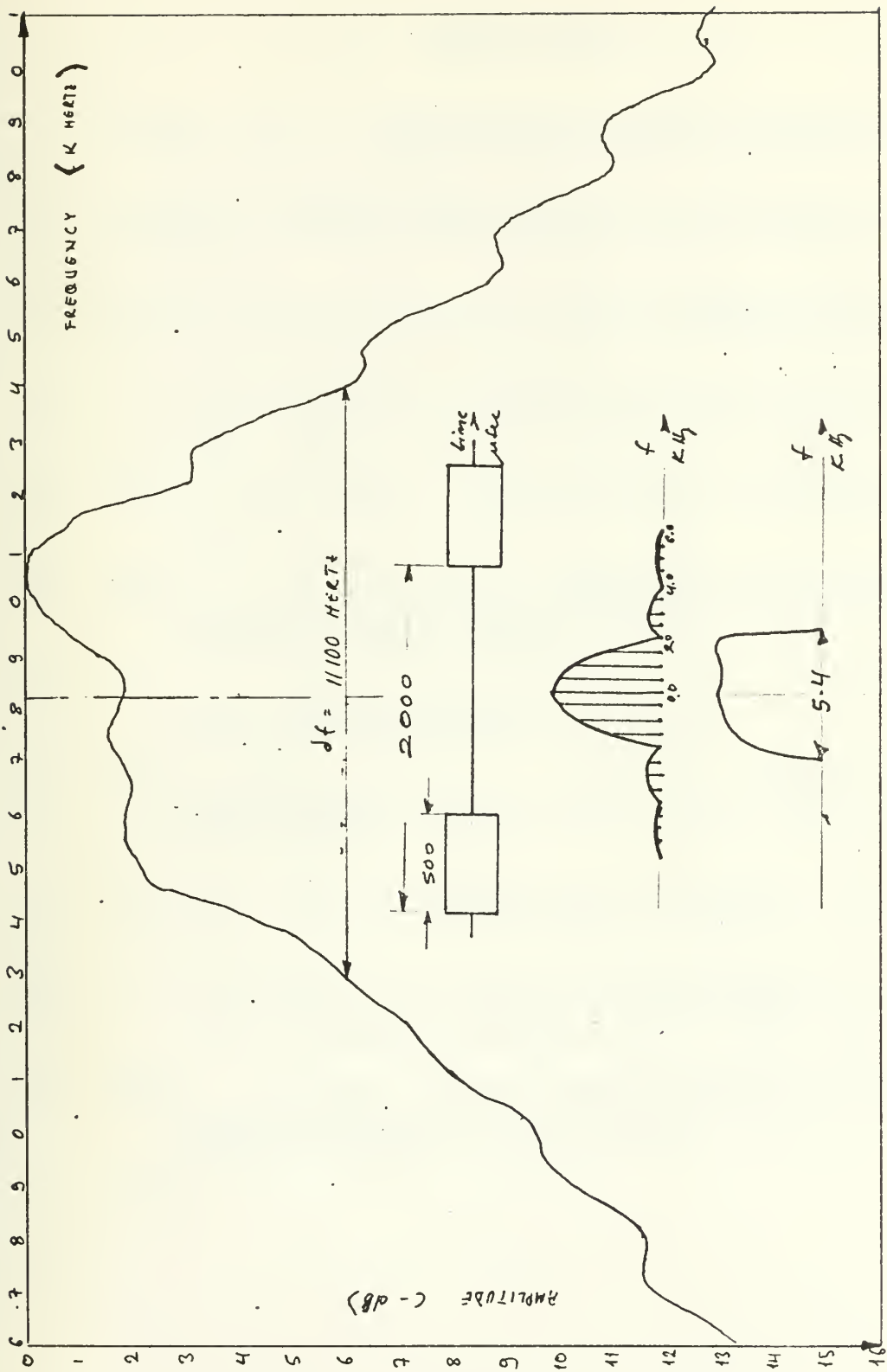
Figure 30 The effect of the width of the input pulse signal on the
IF filter
First sidelobe 3.1 dB lower than the mainlobe
Input pulse width τ = 500 μsec
Duty cycle = 1/4

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